1973

A microwave power amplifier utilizing parallel connected impact diodes

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A MICROWAVE POWER AMPLIFIER UTILIZING PARALLEL CONNECTED IMPATT DIODES

by

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A Thesis
Presented to the Graduate Committee
of Lehigh University
in Candidacy for the Degree of
Master of Science
in
Electrical Engineering

Lehigh University
1973

This thesis is accepted and approved in partial

fulfillment of the requirements for the degree Master of Science.

4/26/73
(date)

Professor in Charge

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ACKNOWLEDGMENTS

The author wishes to express appreciation to Dr. N. Eberhardt of Lehigh University for his guidance and to colleagues at Bell Laboratories for their advice. Dr. C. B. Swan made many technical contributions and reviewed the manuscript. Dr. I Tatsuguchi participated in technical discussions and Dr. J. C. Irvin provided a continuing supply of IMPATT diodes for the experiments. Dr. R. H. Knerr reviewed the manuscript and provided direction and encouragement toward completion of the project.

Finally, the author wishes to thank his wife and children for allowing him the uninterrupted time to complete this manuscript.

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ABSTRACT

Following suggestions by Dr. R. H. Knerr, a broadband FM microwave amplifier using the parallel combination of Schottky barrier IMPATT diodes in a single coaxial resonator has been constructed. In this thesis, after presenting the fundamental theory of operation, the amplifier is described in detail and is compared to six other power combining schemes. The circuit permits the total diode area utilized in a single cavity to be increased significantly beyond that which is practical in a single diode package. This increase is obtained without incurring the thermal dissipation problems associated with parallel diode chips in a single package or series connections of diodes. The parallel diode arrangement allows the use of a single regulated DC power supply without running the risk of thermal runaway of an individual diode. Although each set of diodes used in the amplifier must be matched for similar I-V characteristics, this restriction has not been difficult to meet in practice.

Test circuits were assembled using two and three diodes in parallel and, within the limitations of cavity size and circuit impedance, the scheme should be suitable for extension to four or more diodes. Operating at safe diode junction temperatures, 8 watts output power was obtained (input locking power = 300 mW) at 4 GHz from the two-diode circuit. Under similar conservative operating conditions a three-diode circuit produced 15 watts output at 4 GHz with a locking power of 3.5 watts. The maximum power obtained from the two circuits

was 11 watts and 21 watts respectively. A three-diode circuit, operating at 6 GHz, produced in excess of 10 watts at safe diode temperatures with 2.7 watts input. The measured performance of the three circuits tested demonstrates the feasibility of parallel connecting two or three IMPATT diodes in a single cavity for microwave broadband FM power amplifier applications. The technique has potential for extension to larger numbers of diodes to satisfy applications requiring higher power.

SECTION 1

INTRODUCTION

Solid state, negative-resistance microwave oscillators and amplifiers may be built using the following three classes of two terminal devices; tunnel diodes, IMPATT diodes, and Gunn effect devices. Of these, the IMPATT (Impact Ionization Avalanche Transit Time) diode is probably the most powerful solid state source of CW microwave energy presently available.

IMPATT operation was predicted by W. T. Read in 1958[1]. Read proposed a semiconductor structure composed of a high resistivity region bounded by low resistance end regions such as the n^+-p-i-p^+ diode shown in Figure 1. When this structure is reverse biased into breakdown, a high field avalanche region exists next to a lower field depletion region in which no avalanching occurs. The combined avalanche build-up time and depletion layer transit time are such that the ac current at the device terminals lags the ac voltage by approximately a half cycle, creating a negative resistance in the microwave frequency range. Within the diode's negative resistance region, a portion of the D.C. bias applied to the device terminals may be converted to useful microwave power by properly matching the device to a load.

IMPATT operation was first reported by Johnston, DeLoach, and Cohen in 1965[2]. Eighty milliwatts was obtained at 12 GHz from a simple p-n junction. Misawa has since shown[3] that IMPATT operation is possible in junction diodes with almost any doping profile.

Practical IMPATT diodes are limited in power handling capability by their relatively low efficiency and by heat dissipation requirements. Maximum efficiency in these diodes is achieved at very high current densities. The associated high power densities restrict the maximum area a diode may have and still operate over its entire volume at a temperature low enough to guarantee good reliability[4]. Several solutions to the problem of maximizing efficiency and power output of IMPATT diode structures have been proposed. Among these are the composite diode structure[4], the ring structure[5], and the use of power combining circuits with multiple diodes. The ring structure, although a possible improvement over the commonly used dot structure, might be more difficult to manufacture and still would require power combining techniques to obtain very high power. The composite diode arrangement (multiple small-area chips in a single package) may exhibit moding problems when applied to GaAs devices on a large enough package to permit high power operation. Also, such an arrangement would have required a package redesign which was beyond the scope of this development effort. The third alternative, power combining, is the subject of this investigation.

A number of combining schemes have been advanced. The parallel diode arrangement, proposed herein for transmitter amplifier use in FM radio relay applications, was selected to meet requirements of economy, circuit simplicity, low Q for wide locking range, noise low enough to be compatible with system requirements, and a minimum continuous power output of five watts. Additionally, the parallel diode circuit should be free of parasitic and parametric oscillations. Although the circuit detailed in this paper utilized two diodes, the technique has been successfully applied to three diodes and might be useful with more. The limiting factors in such an arrangement are the ability to tune the circuit with the total capacitance presented by the multiplicity of diodes, the ability to control circuit losses at the low impedance levels of the combined diodes, and the purely mechanical limit to the number of diode packages which can be fitted into the cavity.

SECTION 2

LOCKED OSCILLATOR THEORY

A classic approach to the theory of locked oscillators is described. The theory is then modified to apply specifically to locked negative-resistance oscillators.

2.1 Basic Approach to Locked Oscillators

A simple analysis of locked oscillator operation was presented in 1946 by R. Adler[6]. Although he specifically analyzed a triode vacuum tube amplifier with positive feedback, Adler's treatment is generally applicable to most locked oscillators. Utilizing the plate tuned circuit and definition of terms given in Figure 2, he obtained equations predicting the oscillator behavior under steady state locked conditions, during the transient "pull-in" process immediately after application of a locking signal, and under the influence of a signal outside of the locking range of the oscillator. Only the steady state locked case is described here.

A phasor diagram depicting the relationships between the grid circuit voltages is given in Figure 2. It can be seen from the circuit that $E_g = E + E_1$. α is the angle between the locking signal E_1 and the voltage fed back from the plate circuit E. The total grid voltage is E_g. ϕ, the angle between E and E_g, may be related to α as follows. If E_1 is constrained to be much smaller than E, then $\alpha' \approx \alpha$. Since $\sin\phi = a/E$ and $\sin\alpha \approx \sin\alpha' = a/E_1$, then

$$\sin\phi \approx \frac{E_1}{E}\sin\alpha. \qquad (1)$$

For a single tuned tank circuit as shown

$$\tan \phi = 2Q \frac{\omega-\omega_o}{\omega_o} \tag{2}$$

where Q is the loaded quality factor of the RLC circuit, ω is the frequency of the locked oscillator, and ω_o is the free-running frequency of the oscillator with $E_1 = 0$. Since $E_1 << E$, as stated before, ϕ is small and the approximation $\tan \phi \approx \sin \phi$ permits equations 1 and 2 to be combined to give

$$\sin \alpha = 2Q \frac{E}{E_1} \frac{\Delta\omega_o}{\omega_o} \tag{3}$$

where = has been substituted for $\approx$ and $\Delta\omega_o = \omega_o - \omega$. Equation 3 relates the phase angle α between the locking signal and oscillator signal to the difference between the locked and free-running frequencies. The limitation that $-1 \leq \sin \alpha \leq +1$ defines the locking limits of the oscillator. For values of the right hand side of equation 3 outside this range, the oscillator will be "out of lock". Thus for $\sin \alpha = 1$, equation 3 becomes

$$|2\Delta\omega_{o_{max.}}| = \frac{\omega_o}{Q} \frac{E_1}{E} \tag{4}$$

which gives the total locking range of the oscillator. In FM amplifier applications, such as the one described in this thesis, the peak to peak frequency deviation $2\Delta\omega_o$ is usually maintained at a small percentage of the total locking bandwidth of the amplifier. This minimizes the variation in α and associated phase distortion.

2.2 Negative Resistance Amplifiers and Oscillators

IMPATT diodes are one example of a class of negative resistance devices which can be used as amplifiers and locked oscillators. A few highlights of the operation of such devices, as described by Hines[7] and others preceeding him, will be presented here.

A circulator coupled reflection amplifier is shown in Figure 3a. The components indicated are a circulator to separate the incoming and outgoing signals; a transformer to control the load impedance presented to the diode; a capacitor representing the total capacitance of the diode, diode package, and circuit; a tuning inductor, which includes the inductance of the diode package; and the diode itself. The equivalent circuit of such an amplifier, as seen at the diode, is shown in Figure 3b. G_L is the transformed circulator impedance, i_g the transformed value of the driving signal current, and C and L are as defined for Figure 3a. The fundamental current $I_1(V_1)$ is a nonlinear function of V_1 defined by

$$\frac{I_1(V_1)}{V_1} = Y_1 = G_1(V_1) + jB_1(V_1) \tag{5}$$

where G_1 and B_1 are the diode chip conductance and susceptance respectively. Determination of the nonlinear admittance for IMPATT diodes has been covered in the literature[1,8,9,10] and will not be elaborated upon herein. The reflection coefficient

$$\Gamma = \frac{G_L - Y}{G_L + Y} \tag{6}$$

where $\dot{Y} = G_1(V_1) + j\omega C + \frac{1}{j\omega L} + jB_1(V_1)$, gives the phase of the amplifier response and the power gain $A_p = |\Gamma|^2$. With an analysis similar to that used for Adler's triode circuit, the locked behavior of the negative resistance oscillator may be described. The amplifier equivalent circuit is redrawn in Figure 3c to allow definition of the pertinent circuit currents. The nonlinear diode characteristic is approximated by the current i_1 flowing in the negative conductance $-G$. The Kirchoff node equation of Figure 3c is simply $i_g = i + i_1$. By writing the branch equations for i_1 and i (the circuit current),

$$i = V_1(G_L + j\omega C + 1/j\omega L) \tag{7}$$

$$i_1 = V_1(-G) \tag{8}$$

and eliminating V_1 between them, we obtain

$$\frac{i}{i_1} = -\frac{1}{G}\left(G_L + j\omega C - \frac{j}{\omega L}\right)$$

relating the two currents. Since ϕ is the angle of the transfer function relating i and i_1,

$$\tan\phi = \frac{\mathrm{Im}[G_L + j\omega C - j/\omega L]}{\mathrm{Re}[G_L + j\omega C - j/\omega L]}\,. \tag{9}$$

Using an analysis similar to that used for equation 1, with the constraints that $i_g \ll i$, ϕ is small, and $\alpha' \approx \alpha$ we find that

$$\sin\phi = a/i \qquad \text{and} \qquad \sin\alpha' = a/i_g.$$

Now,

$$\sin\phi = \frac{i_g}{i}\sin\alpha' \approx \frac{i_g}{i}\sin\alpha.$$

With ϕ small, $\tan\phi \approx \sin\phi$, so that

$$\sin\alpha = \frac{i}{i_g}\;\frac{\mathrm{Im}[G_L + j\omega C - j/\omega L]}{\mathrm{Re}[G_L + j\omega C - j/\omega L]} \qquad (10)$$

Equation 10 gives the angle between the oscillator and locking signals as a function of frequency and relative signal amplitude. The maximum value sin α can have is again unity. At this point (α=90°) equation 10 becomes

$$\frac{\mathrm{Im}[G_L + j\omega C - j/\omega L]}{\mathrm{Re}[G_L + j\omega C - j/\omega L]} = \frac{i_g}{i} \qquad (11)$$

for the limits of the locking range of the oscillator. The locking range may be obtained as follows:

$$\frac{\mathrm{Im}[G_L + j\omega C - j/\omega L]}{\mathrm{Re}[G_L + j\omega C - j/\omega L]} = \frac{\omega C - \frac{1}{\omega L}}{G_L} = \tan\phi \qquad (12)$$

Since the resonant frequency of a parallel RLC circuit is given by $\omega_o^2 = 1/LC$, we may substitute $L = 1/\omega_o^2 C$ in equation 12 to get

$$\tan\phi = \frac{\omega C - \frac{\omega_o^2 C}{\omega}}{G_L} = \frac{\omega_o C}{G_L} = \left(\frac{\omega}{\omega_o} - \frac{\omega_o}{\omega}\right)$$

$$= \frac{\omega_o C}{G_L}\left(\frac{\omega^2 - \omega_o^2}{\omega\omega_o}\right) = \frac{\omega_o C}{G_L}\;\frac{(\omega-\omega_o)\;(\omega+\omega_o)}{\omega\omega_o}$$

Now substituting $Q_L = \omega_o C/G_L$ (often called the external Q) and remembering that since ϕ and thus $\tan \phi$ are small so that $\frac{\omega+\omega_o}{\omega} \approx \frac{2\omega}{\omega} = 2$, we obtain

$$\tan \phi \approx 2Q \frac{\omega-\omega_o}{\omega_o} = \frac{i_g}{i}$$

Finally, this may be rearranged to give

$$\left| \frac{\Delta\omega}{\omega_o} \right| = \frac{1}{2Q} \frac{i_g}{i} \tag{13}$$

where $\Delta\omega = \omega - \omega_o$. Equation 13 is simply the dual of equation 4 obtained by Adler for a triode oscillator. Since $i_g = \sqrt{8P_i G_L}$ and $i \approx \sqrt{2P_o G_L}$, the locking bandwidth in terms of the input power and output power (P_i and P_o respectively) of the locked oscillator is

$$2\Delta\omega = \frac{2\omega_o}{Q_L} \sqrt{P_i/P_o}. \tag{14}$$

From this equation it can be seen that the locking bandwidth is inversely related to both amplifier gain and external loaded circuit Q.

SECTION 3

IMPATT DIODE CHARACTERISTICS

The design of a negative resistance amplifier requires a knowledge of the characteristics of the negative resistance device to be used. The following information, although generally true of most IMPATT diodes, is primarily concerned with the type of device used in the proposed power combining scheme.

3.1 Electrical Characteristics

Representative measurements[10] of chip negative conductance $-G_1(V)$ and chip susceptance $B_1(V)$ for a germanium IMPATT diode at particular constant values of bias current and operating frequency are given in Figure 4. Also shown on this plot are the chip power output ($P_o = G_1(V)V^2/2$) and representative conductance lines G_L for both stable amplifier and free-running oscillator operation. For $G_L + G_1(V)>0$, as is always the case with G_L=6 mmhos for the diode measured, the circuit will behave as a stable amplifier with gain and phase given by the reflection coefficient as described previously. If the load conductance is decreased to the point where $G_L + G_1(V)<0$, the circuit will oscillate, without any RF drive, at the RF amplitude determined by the intersection of the load line and the chip conductance. The frequency of oscillation will be that at which the imaginary portion of Y disappears, i.e. $\frac{1}{j\omega L} + j\omega C + jB_1(V) = 0$. The case shown, (with G_L = 2.65 mmhos) is the optimum load conductance for maximum power output from the diode. If the oscillator is operated in the region to the right of the maximum power point, i.e. $G_L <$ 2.65 mmhos, the circuit is undercoupled and any RF input power applied to the diode will merely drive it into a lower power output. In the region to the left of the maximum power point, where 2.65 mmhos $<G_L \leq 5.5$ mmhos, the oscillator is overcoupled. In this region, application of RF drive will increase the diode power output until, with higher drive, saturation is reached at the maximum power point. Thus, for an overcoupled amplifier, it can be seen that the value of load conductance required to achieve maximum diode power output is inversely related to the RF drive level. It has also been shown[11] that, for a given power level, the FM noise is considerably lower when the oscillator is operated in the overcoupled (high load conductance) condition. It should be noted that, for the case shown in Figure 4, where the magnitude of the susceptance is much larger than that of the negative conductance, the real part of the chip admittance is proportional to the real part of the chip impedance.

Measurements performed on the proposed parallel diode amplifier, operated both as a stable amplifier and as a locked oscillator, demonstrated close agreement between the characteristics outlined above and the actual behavior of the amplifier. The physical description of the amplifier, presented in section 5.2, will elaborate on the method used to control the real and imaginary parts of the admittance presented to the diode.

The preceeding relationships between negative conductance, susceptance, power output, and the RF chip voltage permitted easy visualization of the negative resistance amplifier behavior of the device. Parameter variations as a function of frequency, bias current, and temperature will be considered next.

The admittance versus frequency behavior of an IMPATT diode may be predicted by the use of a parallel RLC equivalent circuit in which the resistance and inductance represent the admittance of the avalanche region and the capacitance is that of the depletion layer of the reverse-biased diode. For the Read diode[1], described in the introduction, the lowest frequency for which the real part of the chip admittance can become negative is equal to the resonant frequency of the parallel circuit. For a general IMPATT diode, (and in particular for the diodes used in the proposed amplifier) the cutoff frequency, below which the conductance can no longer become negative, may be well below the resonant frequency of the chip and is proportional to the relative width of the avalanche and drift regions. The admittance of a typical IMPATT diode is directly proportional to bias current whereas the inductance is inversely proportional to the current. Since the depletion layer capacitance is relatively independent of diode current, the resonant frequency increases approximately as the square root of the bias current. Transforming the parallel RLC circuit described above to the series equivalent and considering the impedance of the device, it can be seen that the chip will exhibit capacitive reactance below resonance and inductive reactance above. In the amplifier proposed in this thesis, the diodes were operated below resonance and were tuned with a series inductance.

The small signal Q of the diode, which controls the threshold and build-up rate of oscillations of the device, (when operated as a negative resistance oscillator) is defined as the time average of the AC field energy divided by the AC power dissipation per cycle[12]. For amplifiers, low Q is desirable, a situation which improves with increasing bias current for a given operating frequency. For a fixed bias current, a typical IMPATT structure has an optimum frequency, defined as the frequency at which the magnitude of the small signal Q is a minimum.

3.2 Thermal Considerations

Thermal considerations are important to an IMPATT amplifier design because of the high current densities required for maximization of diode efficiency. In order to achieve high gain and high power output, IMPATT diodes are usually operated near the maximum temperature consistant with reasonable diode reliability. Using carefully designed packages coupled with diamond heat spreaders[13], experimental diodes have been fabricated which have produced in excess of eight watts output at 4 GHz. For the amplifier proposed in this thesis, the behavior of breakdown voltage versus diode junction temperature was a critical parameter. The reverse breakdown of a p-n junction is dependent upon the ionization rates of holes and electrons. Since these rates decrease with rising temperature[14], the breakdown voltage of an IMPATT diode is directly proportional to temperature. Thus, the power sharing between paralleled diodes in an amplifier, utilizing a common bias feed for all diodes, will be thermally stable and runaway of a single diode will not occur.*

3.3 Power-Impedance Limitation

A basic limitation, beyond the thermal restrictions on IMPATT diode power output, is the power-impedance product put forth by DeLoach[15]. The maximum RF power (P_m) a solid state device can produce is limited by the critical electric field (E_c), at which avalanche breakdown occurs, and the scattering limited velocity (v_{sl}) obtainable by carriers within the semiconductor. If C is defined as the effective parallel plate capacity of the semiconductor device, when it is swept free of mobile charges, and X_c is the reactance $1/2\pi fC$, then the equation obtained by DeLoach[15]

$$P_m X_c f^2 = \frac{E_c^2 v_{sl}^2}{4\pi^2} \tag{8}$$

gives the relationship between maximum power output and device impedance for a given frequency. The maximum power obtainable from a semiconductor device is thus approximately proportional to $1/f^2$.

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*Freedom from thermal runaway requires that the paralleled diodes be constrained to have the same RF voltage. Runaway effects related to rectification characteristics can occur when the diodes are biased from a common DC source but are RF isolated.

SECTION 4

POWER COMBINING

In the introduction, it was pointed out that it is often desirable to combine the outputs of two or more diodes to obtain high power from amplifiers and phase locked oscillators operating in the microwave region. In this section a number of the power combining schemes, that have been proposed, will be described. These schemes include four utilizing parallel combination of oscillators and two using series connected diodes in a single resonator. Finally, the parallel diode technique proposed in this thesis will be introduced.

4.1 Parallel Combination of Oscillators

Power combination of the outputs of discrete oscillators to achieve higher power levels than can normally be obtained with single oscillators has several advantages over other possible techniques. The impedance levels of such arrangements are high enough to prevent excessive circuit losses and the schemes can accommodate some mismatch of operating breakdown voltage and RF characteristics between diodes. Furthermore, heat sinking of diodes mounted in individual oscillator cavities is relatively easy to accomplish. However, parallel oscillators require individual cavities, transformers, and bias circuits-if not separate bias supplies, so most of the oscillator combining arrangements suffer from poor economy of components, lack of circuit simplicity, and in some cases, the problem of tuning each of the oscillators separately. Four proposed arrangements using parallel combination of oscillators will be described.

4.1.1 C. T. Rucker's Symmetrical Power Combiner

In December of 1969[16], C. T. Rucker described a circuit in which five IMPATT diode oscillators are connected, in a symmetrical array, through quarter wave transformers and individual coupling capacitors to a common output port. The output ends of the five center conductors have built-in stabilizing resistors which are, in turn, connected to a common bias network. A schematic diagram of two sections of the oscillator is shown in Figure 5. The output capacitors, which were physically realized using a single disc on the output side, provide cross coupling between the oscillators to phase lock them to each other. It should be noted that, although half-frequency instabilities and RF voltage inequalities between oscillators will be absorbed in the resistor network, the desired RF output current does not flow through the resistors as long as the oscillator signals have the same amplitude and phase. A disadvantage of the Rucker combiner is that the DC bias current for each diode must flow through the stabilizing resistors resulting in a loss of efficiency. The bias network is simply a high impedance line bypassed with a capacitor approximately one quarter wavelength (at the operating frequency) from the common connection between the oscillators.

An analysis of Rucker's oscillator, published by Kurokawa in November of 1970[17], employed an eigenvector approach to study the circuit's behavior. The analysis verifies that, with appropriate loading conditions, Rucker's scheme should be free of undesired modes of oscillation and should provide the added benefits of improved AM and FM carrier to noise ratio.

Operating the five diode circuit in the free running mode, Rucker obtained about 4.5 watts at frequencies near X-band. A more recent paper[18], suggesting the use of the Rucker circuit as a reflection amplifier or phase-locked oscillator, predicted output powers of 13 watts at 6dB gain using presently available two watt diodes.

However, a limitation may exist on the practical usage of Rucker's combining technique. The oscillator shown uses a single power supply to bias multiple diodes. This arrangement may be undesirable since the individual diodes may have unequal RF voltages and a common DC bias supply could result in thermal runaway as described earlier.

4.1.2 Circulator - Mismatch Power Combiner

A second power combining technique, proposed by L. F. Moose and M. D. Bonfeld, was described by I. Tatsuguchi in December of 1970[19]. The technique is shown schematically for three sections in Figure 6a. The schematic of a single splitter-combiner section (Figure 6b) may be used to describe the operation of the circuit. Although the circuit can be used for both power splitting and power combining, only the latter mode of operation will be described in detail.

The basic combiner section is composed of two interconnected circulators with a mismatch in the common arm which is adjustable in magnitude and position. Two signals of equal frequency but arbitrary magnitudes and phases E_1 and E_2, incident upon the input ports of the circulators, will circulate around to the common arm as shown. Each signal will be partially reflected toward the circulator from which it came and thence circulated to the output port of that circulator. Since the reflection coefficient ρ and the transmission coefficient τ are related by $\rho^2+\tau^2=1$ (barring absorption losses), the remainder of each signal will be transmitted through the mismatch to the output of the opposite circulator. The resultant signal appearing at the output of circulator 1 is given by the equation

$$E=E_1|\rho|e^{-j(\beta L-\phi\pm 2\beta\Delta\ell)}+E_2|\tau|e^{-j(\beta L-\phi+\frac{\pi}{2})}, \tag{9}$$

where L is the path length traversed by either signal from the input of one circulator to the output of the other, and $\Delta\ell$ is the displacement of the mismatch position from the center of the common arm. By proper adjustment of the magnitude and displacement of the mismatch, the two signals can be made to add at the output of one circulator and cancel at the output of the other. This applies only as long as the initial relationships between the amplitudes and phases of the input signals remain unchanged.

The components in Figure 6a are identified as follows. The top row of circulators and their associated mismatches serve as power splitters for the input locking signal. The second row of circulators are isolators used to separate the diode oscillators from the input and to absorb any uncombined power from the combiners. The remaining two circulators and associated mismatches are the power combiners, with the output issuing from the circulator on the lower right. Each oscillator is individually biased and each is individually tuned with an adjustable quarter wave coaxial transformer. Details of each coaxial oscillator circuit are identical to those shown in Figure 12, the single diode housing used for the testing of individual diodes for this thesis. Any number of modules can be combined in the manner described; however, each additional module adds a circulator pass to the path a signal must follow through the combiner, so circulator losses become a limiting factor on the number of combining stages that may be used. In addition, the multiplicity of circulators and stringent requirements on the magnitude and phase relationships of the signals from each diode may result in excessive complication of amplifier tuning. The results reported in 1970 for the three module circuit are as follows: Nearly two watts was obtained from three silicon diodes operating with a locking gain of 16.5 dB and an FM noise measure of 47dB. This noise figure was subsequently improved by 6 to 10 dB through circuit optimization.

4.1.3 A Single Cavity Waveguide Combiner

In January of 1971, K. Kurokawa and F. M. Magalhaes reported on the successful power combination of twelve packaged IMPATT diodes coupled to a single waveguide cavity[20]. A further, thorough mathematical analysis of the oscillator was published in October 1971[21]. A cutaway drawing depicting the mechanical structure of four sections of the oscillator is shown in Figure 7. Each diode is mounted in a coaxial housing and is coupled to the waveguide cavity through a coaxial transformer. The DC bias is supplied to each diode by the center conductor which extends through a tapered termination mounted in the opposite end of the coaxial housing from the diode. In experiments, the termination consumed about three percent of the generated power of the diodes[22], considerably less than the loss of the cavity and diode series resistance. However, this oscillator circuit had a high external Q and reducing the Q increases the loss in the termination. For this reason, the waveguide combiner is not suitable for use in broadband FM amplifier applications. Again using an eigenfunction approach, Kurokawa demonstrated that this circuit is free from moding problems and will therefore give stable, single-frequency operation. Reported performance of a twelve diode oscillator was 10.5 watts at 9.1 GHz[20].

4.1.4 Resistive Hybrid Combiner

The resistive hybrid has been used as a power combining and splitting circuit for over a decade Within a year after the first successful demonstration of IMPATT diode operation, H. Fukui reported on a power combining circuit which utilized multiple hybrid combiners to increase the power output from an oscillator using IMPATT diodes[23]. Fukui's combiner circuit is shown along with a sketch of a single microstrip hybrid combiner[24] in Figure 8. The operation of the combiner is such that two signals of equal phase and magnitude applied to the side ports of the hybrid combine to yield double the power at the output. Differences in phase or magnitude will result in power loss in the resistor. Also, since half of the power from uncorrelated noise appearing at the side ports will be dissipated in the resistor, the hybrid can provide a 3dB improvement in the signal to noise ratio after combination.

Using eight diodes in the combining scheme shown, Fukui successfully demonstrated power addition at 6.1 GHz with about 7 dB improvement in single sideband carrier to noise ratio over the averaged value of eight individual oscillators. The locking bandwidth of the eight diode circuit was approximately 50 MHz at 10 dB gain. The narrow bandwidth of the circuit is attributed to the high Q (nearly 100) of the individual oscillator circuits. Coaxial oscillators of the type shown in Figure 12 could be used in this power combiner with greatly improved locking bandwidth. The number of oscillators which can be combined using the hybrid circuit is unlimited in theory; however, the hybrid combiners are not lossless and overall circuit losses become significant when many levels of combining are used.

4.2 Series Combination of Diodes in a Single Resonator

Series connection of IMPATT diodes to increase the microwave power available from an amplifier or oscillator has a large potential advantage over similar parallel schemes. The increase in output power is not accompanied by a decrease in impedance level, thus the series connection provides a way around the $1/f^2$ power impedance product limitation of IMPATT diodes. The disadvantage of using the series connection lies in the difficulty of providing an efficient heat sink. Two series connected diode arrangements which at least partially circumvent the heat sinking problem are presented here.

4.2.1 Packaged Diodes in Series

In February of 1968, F. M. Magalhaes and W. O. Schlosser reported on an oscillator using three packaged IMPATT diodes in series as shown in Figure 9[25]. The copper housing formed a coaxial cavity which was tuned with a triple slug tuner. The heat path through the Beryllia rings was calculated to be only 0.8°C/watt higher than the normal path through the backs of the diodes. The results obtained from the experimental setup verified that power addition of the individual diode outputs occurred as expected. It should be noted, however, that for diode DC power inputs in the order of thirty watts or more as required for the amplifier proposed in this thesis, the 0.8°C/watt thermal resistance increase of the BeO rings is not negligible. Also, the tolerances required to maintain good thermal contact between the diodes and rings and between the rings and the copper housing might be difficult to obtain in routine manufacture.

4.2.2 Series Connected Chips in a Single Package

A second approach to power addition from series connected IMPATT diodes was suggested by J. G. Josenhans[26]. This scheme consisted of bonding three diode chips on separate metallized pads on a single diamond heat sink and stitch bonding the three chips in series. As was the case for the series arrangement presented in 3.3.1, power addition was observed experimentally and FM oscillator carrier to noise ratio was improved by a factor of three as a consequence of the increased power output. Although a power output of nearly five watts was obtained by Josenhans, the capability of the arrangement is still limited by the power than can be dissipated from a single diode package and the maximum distance that can be tolerated between chips.

4.3 A Parallel Diode Amplifier

In the preceeding paragraphs, a brief survey of available techniques for combining the power outputs of multiple oscillators and diodes has been presented. Each of the techniques described suffers from one or more of the following disadvantages:

1. Multiplicity of oscillator components, bias circuits, and power supplies.
2. Separate tuning of the individual oscillators in a combining circuit.
3. Difficulty of removing the heat generated by multiple diodes or multiple chips.
4. Excessively high Q, preventing oscillator usage as a broadband FM amplifier.

A microwave amplifier, designed to avoid the difficulties listed above, utilizes parallel IMPATT diodes mounted in a single coaxial oscillator cavity, tuned and matched with a single transformer, and biased from a single bias circuit and power supply. This circuit is described in detail in Section 4.

SECTION 5

MICROWAVE AMPLIFIER USING PARALLEL IMPATT DIODES

A simple microwave amplifier, utilizing the parallel combination of packaged IMPATT diodes in a single resonator, is presented. The section begins with a summary of the requirements controlling the design of the amplifier. The summary is followed by a physical description of the circuit and a discussion of physical and electrical characteristics of the Gallium Arsenide diodes used in the amplifier. Next, the more significant measurements performed on the test amplifier are listed and discussed. Finally, the results obtained from an experimental 3-diode circuit are offered as proof that the two diode circuit extensively tested may indeed be successfully extended to three or more diodes.

5.1 Design Requirements

An application exists for a solid state amplifier which provides five to ten watts output in the common carrier band at 3.7 to 4.2 GHz. The amplifier must be sufficiently inexpensive to compete with existing "tube-type" amplifiers for both replacement and new equipment usage. Since the power combining scheme to be discussed in this thesis involves only the final stage of the required amplifier, the design guidelines which follow are limited to that stage wherever possible.

The amplifier output stage is required to furnish a minimum of 37 dBm output when driven by an input signal of 25 dBm. Operating at this level, the combined noise figure of the entire amplifier as calculated from Friss's formula must be consistant with the requirements of the microwave radio system. The amplifier must be capable of handling a frequency modulated signal with a certain peak deviation without excessive intermodulation distortion as determined by noise loading tests. Finally, the operating conditions of the amplifier must be such that the amplifier exhibits the same high reliability expected from other solid state amplifiers.

5.2 Physical Description of Amplifier

The schematic diagram of a parallel IMPATT diode phase-locked oscillator, designed to fulfill the requirements listed above, is shown in Figure 10. For purposes of discussion, the amplifier is divided into the three parts indicated; a coaxial oscillator assembly, a stripline bias block and circulator, and a bias circuit.

5.2.1 Coaxial Oscillator Assembly

The twin diode coaxial oscillator assembly is diagrammed in Figure 11. For comparison, Figure 12 is a diagram of the single diode coaxial housing. A coaxial circuit was chosen for low Q and thus wide locking bandwidth. A pair of IMPATT diodes is firmly held in an OFHC copper mount by threaded OHFC copper plugs which provide a low thermal impedance path from the diode stud to the mount. The diode mount forms the top wall of the oscillator cavity and serves to conduct the large amount of heat generated in the diodes to a finned heat sink (not shown). The close proximity of the diode center contacts to each other permits short connections to the coaxial center conductor, thereby reducing the series inductance of the oscillator circuit and enabling the tuning of high capacity (large area-high power) diodes. The coaxial transformer provides an impedance match between the 50 ohm coaxial line and the parallel diodes, which in normal operation, have a combined AC resistance of several ohms. Adjustment of the load resistance presented to the diodes is achieved by changing transformers and thereby changing the transformation ratio. It should be noted that the resistance shown on the power-noise and power-versus current curves described in section 4.5 is the calculated characteristic impedance of the quarter-wave section which serves as a transformer. The actual resistance seen by the oscillator or amplifier cavity is given by

$$R_D = \frac{R_T^2}{50 \text{ ohms}} \tag{15}$$

where R_D is the resistance presented to the cavity and R_T is the characteristic impedance of the transformer. The axial spacing between the transformer and the diodes is adjustable, to facilitate frequency tuning of the oscillator by varying the reactance presented to the diode.

The "fuzz buttons" served as low resistance, low inductance contacts between the diode center terminals and the center conductor of the coaxial housing. Recent experiments have been performed with short, gold plated beryllium copper, spring contacts in place of the "fuzz buttons" shown. This change resulted in improved contact reliability and greater ease of circuit assembly, but was not introduced until after all of the data shown for the two diode circuit had been obtained.

The idea of operating IMPATT diodes in parallel, to achieve higher power than is possible from a single diode, is not a new one. However, as pointed out in the literature,[17,20,26] a moding problem can exist, in which the paralleled diodes oscillate in push-pull. Indeed, the first attempt at paralleling diodes, which used a solid center conductor to contact the diodes, produced an assymetrical oscillation at approximately 5.1 GHz. This oscillation was only weakly coupled through the circulator to the output and was therefore not immediately discovered, or recognized as the cause of the amplifier's failure to exhibit efficient power combination. When the nature of the undesired oscillation was perceived, it was suggested[27] that a quarter-wave trap, terminated by a resistor, be used to extinguish the assymetrical mode. The suggested damping technique, realized by the teflon filled slot with a 22 ohm chip resistor connected across the end, has been completely successful in eliminating the push-pull modes of oscillation. Subsequent work with amplifiers utilizing the terminated slots has indicated that, although the slot was dimensioned to be a quarter-wavelength at 5.1 GHz, the length of the slot is noncritical. The slot shown has been effective in eliminating assymetrical mode oscillations occurring over a twenty percent band of frequencies. An important feature of the damping arrangement used is that, for a well matched pair of diodes operating in the desired mode, no DC current and negligible RF current flows through the resistor. For the push-pull mode, the quarter-wave slot appears as an open circuit, thus the opposing RF voltages on the diodes would appear across the resistor. This situation causes such modes to be heavily loaded and is the source of their elimination.

5.2.2 Circulator and Bias Trap

Signal paths for RF input and output and the D.C. bias connection were provided on a suspended ceramic substrate transmission-line circuit. This circuit, shown schematically in the center of Figure 10, was connected to the coaxial diode housing through a flexible bellows. The circulator served to separate input and output signals, the capacitor (an overlay type) isolated the RF ports from the high voltage required for the diodes, and the bias trap provided more than 40 dB isolation between the RF circuits and the external bias circuitry. The bias trap was composed of two quarter-wave sections of high impedance line and two quarter-wave 50 ohm open circuited stubs arranged as a two section band reject filter. The connection from the bias trap to the remainder of the bias circuit was made with a short length of wire through several ferrite beads and a BNC connector.

5.2.3 Bias Circuit

At the time the experimental work on a 4 GHz two parallel diode circuit was taking place, the requirements for a bias circuit, capable of preventing parametric oscillations and tuning induced burnouts, were believed to be fairly simple. What was desired was a bias supply capable of behaving as a current regulated source up to the region of several hundred megaHertz or higher. Since commercially available current regulated power supplies and their associated cabling have high output capacitance, the function of the bias circuit was to extend the high impedance characteristics of a current regulator to the desired frequency, without incurring a high voltage drop in the circuit. A circuit designed to meet these requirements was used in the amplifier. Voltage and current monitoring features were also incorporated into the bias circuit in the manner shown. Recent work on controlling parametric oscillations and tuning induced burnouts[28] has indicated the desirability of presenting an impedance to the diode which is low at the subharmonic of the operating frequency and high but well controlled at frequencies below the subharmonic. As features to implement such a characteristic were not used in the 4 GHz amplifier, they are not shown in the circuit.

5.3 Diode Characteristics

The success of the parallel diode amplifier design project was completely dependent upon the availability of matched sets of high power diodes. The devices used in the experiments reported here were noncommercial Gallium Arsenide Schottky barrier IMPATT diodes, each of which was capable of approximately four watts output under safe operating conditions as an oscillator.

The physical structure of the diode chip is shown in Figure 13[29]. The active region is the epitaxial layer, which has an impurity concentration of roughly $6 \times 10^{15} cm^{-3}$. The platinum-gold metallizations provide low resistance contact to both ends of the diode as well as forming the Schottky barrier junction with the epitaxial layer. The active side of the diode chip is thermal compression bonded to a metallized diamond which spreads the heat into a copper stud forming the base of the diode package. An alumina cylinder, bonded to the stud, serves as an insulating housing for the diode chip. A copper cap on the cylinder completes the package and provides the contact to the ungrounded side of the diode. The thermal conductance of the diode structure is such that a DC power input of thirty-two watts is considered safe when the package is properly mounted in a good heat sink.

The series resonance of the diode assembly was above the operating frequency of the experimental amplifier, thus the diodes exhibited capacitive reactance and had to be tuned with a series inductance. The large-signal negative conductance of the diodes was such that maximum efficiency was obtained with a resistance of approximately five ohms presented to each diode by the transformer.

Ideally, for optimum power addition in the proposed circuit, the breakdown voltages at normal operating current and the large-signal susceptances and negative conductances should be well matched. In practice, no great effort was made to match the conductances and susceptances for the experiments. In general, the diodes used came from the same diffusion lot and were grouped for an operating breakdown voltage mismatch of no more than one volt. It was found that, for diodes from the same lot, a ± 1 volt tolerance on breakdown match was not too difficult to meet with reasonable care in grouping.

5.4 Experimental Results

In order to determine the suitability of the proposed power combining scheme, a number of measurements were performed on the test circuit described in section 4.2. The tests performed provided data on the following: 1. Power output as a function of current, frequency, and transformer impedance; 2. FM single sideband noise as a function of output power and transformer impedance; and 3. Locking range as a function of locking power. To provide a basis for comparison, these measurements were performed on both the individual diodes and the pair. Although a number of diodes have been tested in the circuit, the results presented were obtained using the pair for which data is listed in the table below. Both the diodes and data were supplied by Bell Laboratories at Murray Hill.

Table of Diode Characteristics

Parameter	Diode #1	Diode #2
Breakdown Voltage (V_B)		
at 10µA	115 V	115 V
1mA	118 V	116 V
10mA	120 V	120 V
Zero Bias Capacitance C_o:	18.8 pF	19.2 pF
Breakdown Capacitance C_B:	1.5 pF	1.6 pF

To determine the transformer impedance required for optimum power output, a series of power versus current curves was generated for each diode and for the pair. Representative curves for the better impedance values are shown in Figure 14 through 21. Power curves were made for 3.7 GHz, 3.95 GHz, and 4.2 GHz, the bottom, middle, and top, respectively, of the frequency band of interest. The single diode curves demonstrate very close concurrence of the diode characteristics, and exhibit very little power difference between the 16.0 ohm and 18.1 ohm transformers. Since the larger impedance transformer causes a reduction in the FM noise, as will be shown later, it is probably the best choice for single diode operation.

The power curves for the diode pair show a slight drop in power for a change in transformer impedance from 9.2 ohms to 10.6 ohms, and a more significant drop as the impedance is increased to 12.7 ohms. An attempt to determine the maximum power output available from the pair, using the 10.6 ohm transformer (which proved to be more parametrically stable at high powers), is shown in Figure 21. The maximum efficiency, (defined as $\frac{\text{RF power added}}{\text{DC power input}}$) of the diode pair was 12.3%. This, when compared to the 13.3% figure obtained for a single diode indicates that power addition was achieved. The actual power combining efficiency is perhaps better determined using power-noise curves.

Noise measurements were made on both individual diodes and on the pair, using the techniques described by Tatsuguchi et al[11]. The results obtained for the single diodes are shown in Figures 22 and 23. These curves are typical of the results obtained for many diodes of this type and point out the slight difference in the noise characteristics of the two diodes used in the combiner. In order to provide a fair basis for comparison with the noise curves of the diode pair, the single diode power-noise characteristics were measured with 150 mW of locking power while 300 mW was used with the pair. Figure 24 is the set of power-noise curves for the diode pair, along with a repeat of those for diode number 1. Comparison of the two sets of curves gives an indication of the quality of power addition obtained from the parallel diode amplifier. At high power, for a given noise figure and power input per diode, points of 3 dB power addition (doubling of power) can be found for the pair as contrasted to the single diode. This comparison is made somewhat difficult, however, by the large effect on noise caused by a small step in transformer impedance.

When an amplifier, of the type considered, is operated in a phase-locked oscillator mode, the locking range becomes important as an indicator of the maximum deviation an FM signal can have and still be amplified with low phase distortion. The experimental circuit described above behaved as an oscillator with the transformer impedances chosen. For this reason, the power frequency curves of the single diode oscillator and the diode pair oscillator are shown in Figures 25 and 26 respectively. The curves clearly demonstrate the improvement in locking bandwidth obtained with increased locking power. For a given power input per diode, the locking bandwidth of the diode pair and the single diode are essentially the same. Using equation (14) and the locking bandwidth with an input power of 600 mW, the external Q's of the single diode oscillator and of the twin are 6.0 and 5.1 respectively. The similarity between the two is to be expected since both oscillators used the same coaxial circuit. The transformer impedances were chosen to provide a similar per-diode load for the locking measurements in each case.

5.5 A Three-Diode Parallel Circuit

Utilizing the same combining scheme described for the two diode circuit, a three diode oscillator has been tested. The three parallel diodes were mounted on 120° centers with the center contacts again in close proximity. A delta connected resistor chip was mounted on top of a three-way slot in the center conductor of the coaxial housing. The remainder of the oscillator circuit was identical to that used for the diode pair.

Figure 27 gives the power versus current curve obtained for the three diode circuit at 2950 MHz with an input locking power of 3.5 watts. Using a 9.1 ohm transformer and a DC input power of approximately forty watts per diode, a power output in excess of twenty watts has been observed. At a safer DC power input of 32 watts per diode, 15 watts output was obtained with otherwise identical conditions.

A similar circuit has produced over ten watts output at 6.2 GHz with an average DC input of 32 watts per diode and a locking power of 2.7 watts. The power vs current curve for this case is shown in Figure 28. Again a 9 ohm transformer was used for the test.

It should be noted that as the number of diodes combined in parallel is increased, the required transformer impedance decreases approximately as $1/\sqrt{n}$ where n is the number of diodes used. The tolerable circuit impedances thus determine the maximum number of diodes which can be combined in this manner.

SECTION 6

CONCLUSIONS

A simple modification to the structure of a single diode coaxial oscillator circuit permits the paralleling of two or more diodes for increased microwave power output. In experiments performed at four and six gigaHertz with both two and three diode arrangements, the amplifier demonstrated efficient power combination with no sacrifice in noise or locking bandwidth. Economies of the proposed technique, as compared to other power combining schemes, accrue from the need for only one coaxial circuit, bias circuit, and current stabilized power supply. Additional savings should be realized from the simplified tuning requirements which are inherent in the single oscillator over those of a number of oscillators combined. The number of diodes which may be combined in the proposed manner is limited primarily by the circuit losses present at the low impedance level of the paralleled diodes. For the two and three diode oscillators tested, the worst case circuit impedances were several ohms or more, thus permitting easy realization of fairly low loss coaxial components.

A possible disadvantage of the proposed circuit lies in the requirement for matched diodes. In order to compensate for unequal division of the bias current among mismatched diodes, it may be necessary to derate the maximum DC power input to assure acceptable reliability.

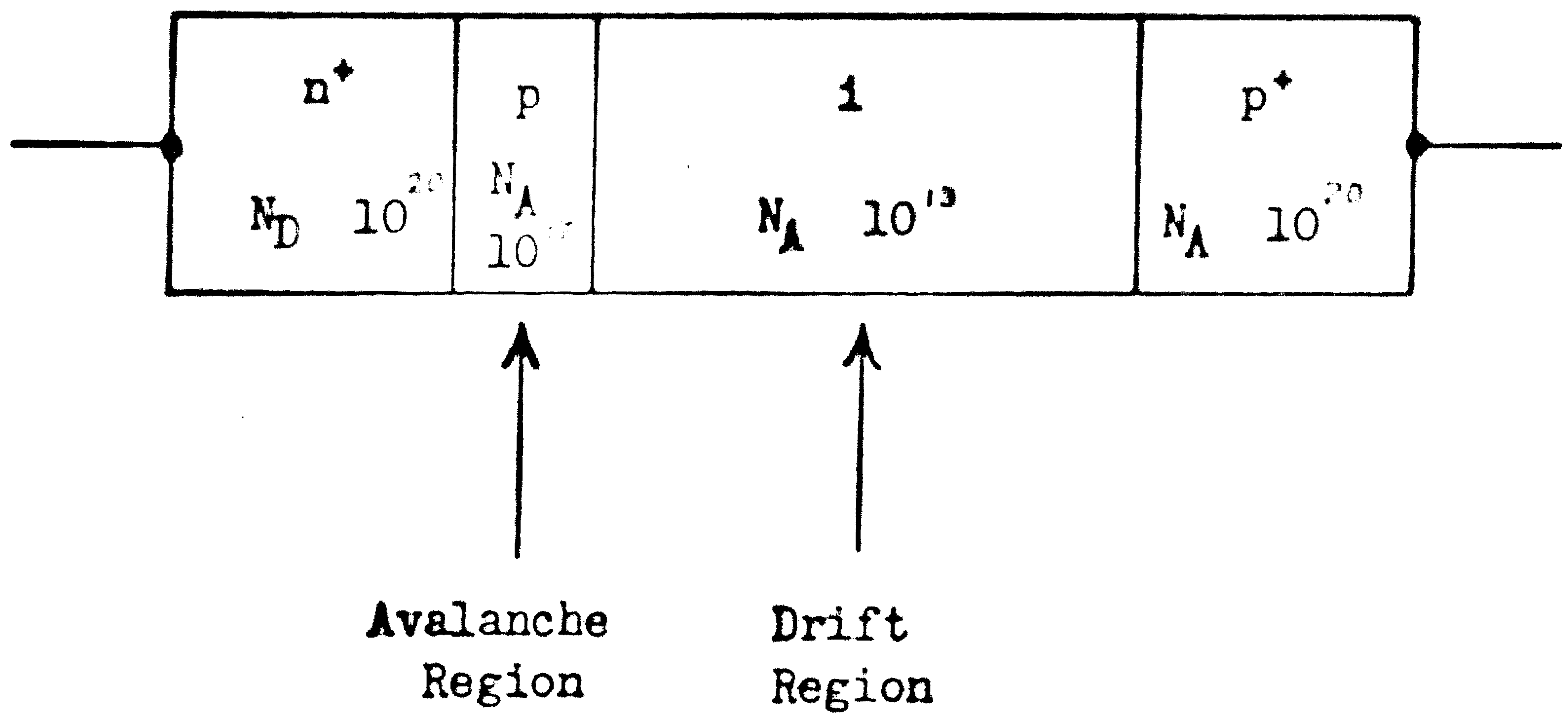


FIGURE 1. STRUCTURE OF THE READ DIODE [1]

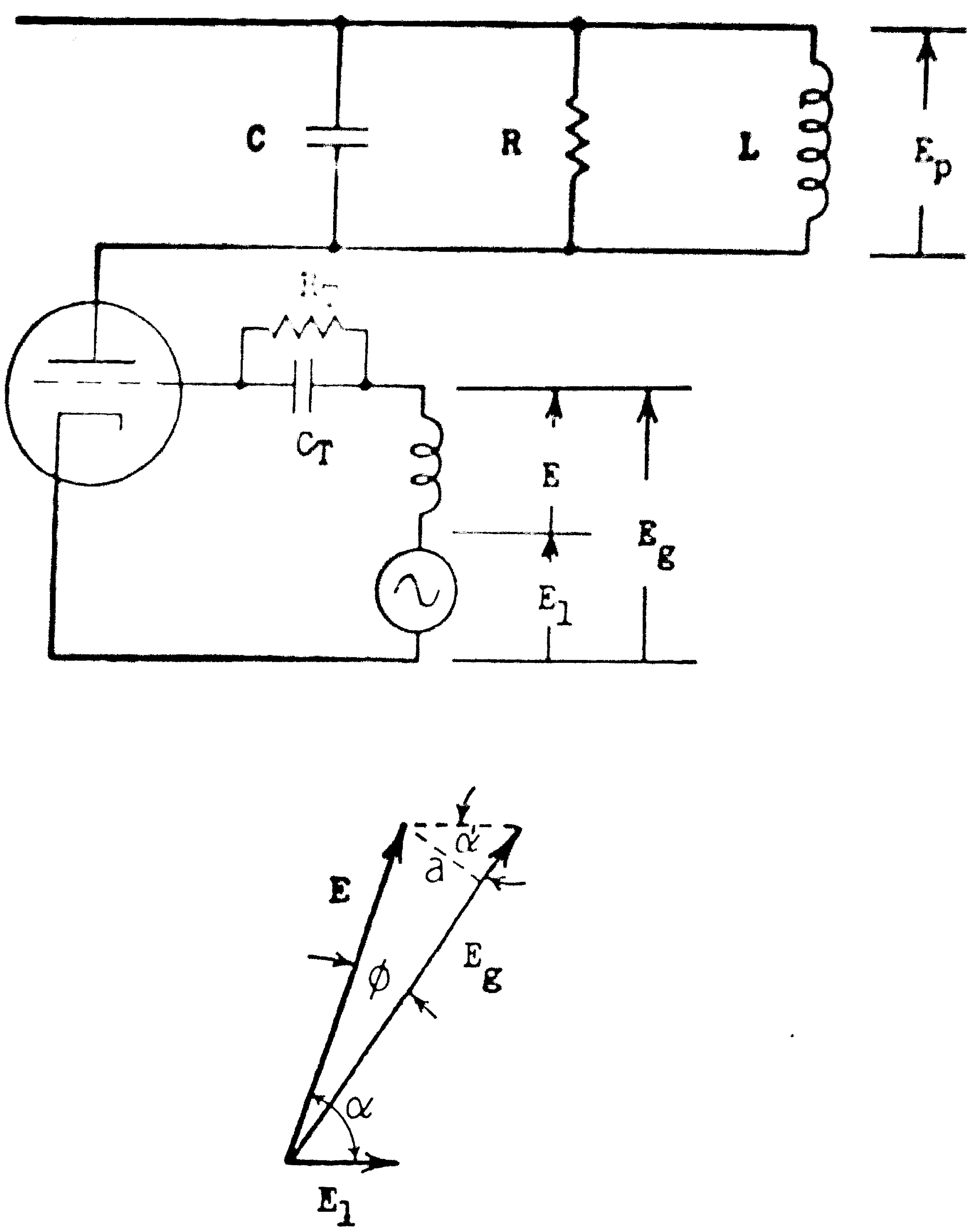


FIGURE 2. PLATE TUNED OSCILLATOR CIRCUIT OF R. ADLER[6]

E_p = Voltage across Plate Load

E = Voltage induced in Grid Coil

E_1 = Voltage of Locking or Input Signal

E_g = Resultant Grid Voltage

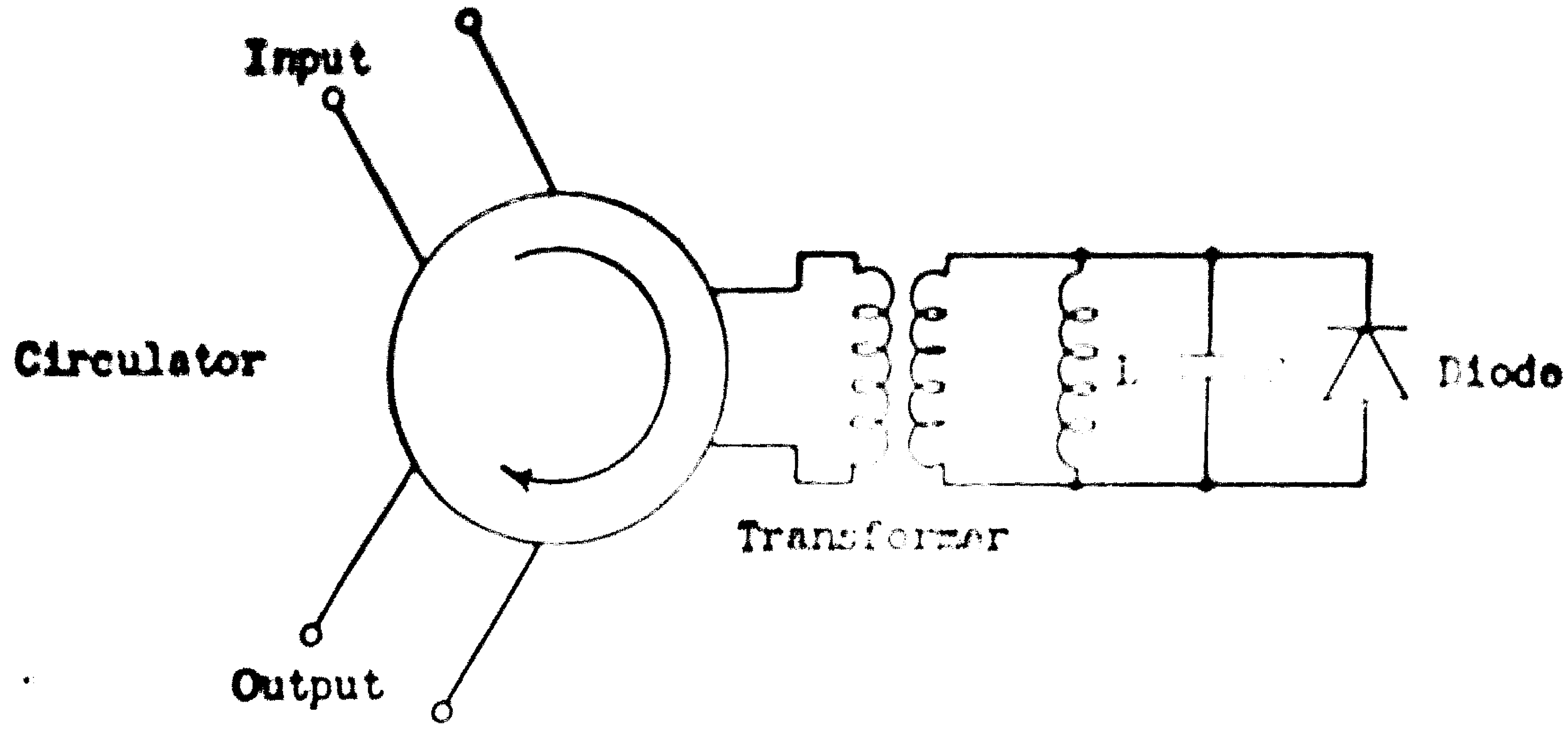


FIGURE 3a. REFLECTION AMPLIFIER

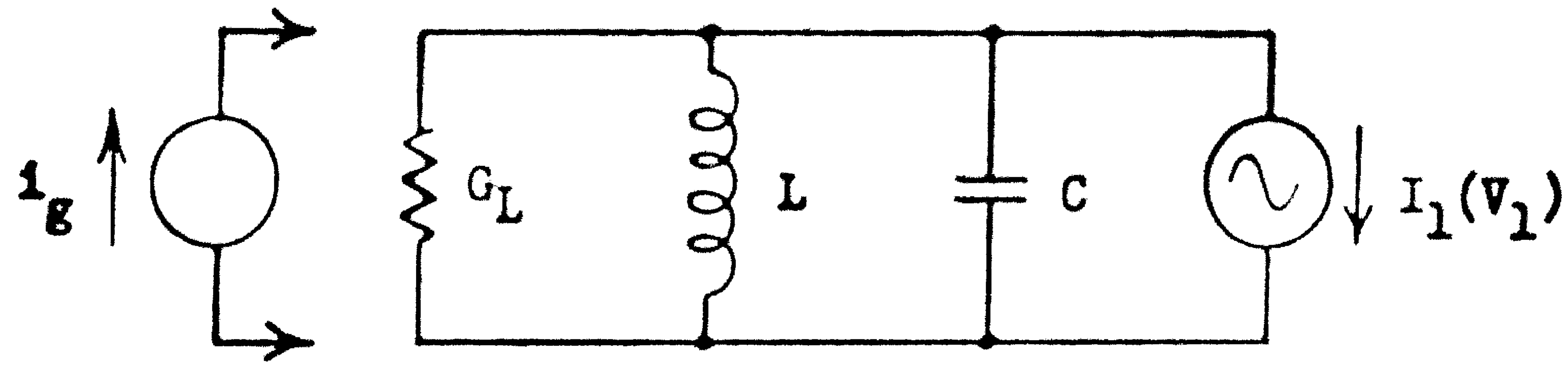


FIGURE 3b. EQUIVALENT CIRCUIT OF REFLECTION AMPLIFIER AT DIODE[7]

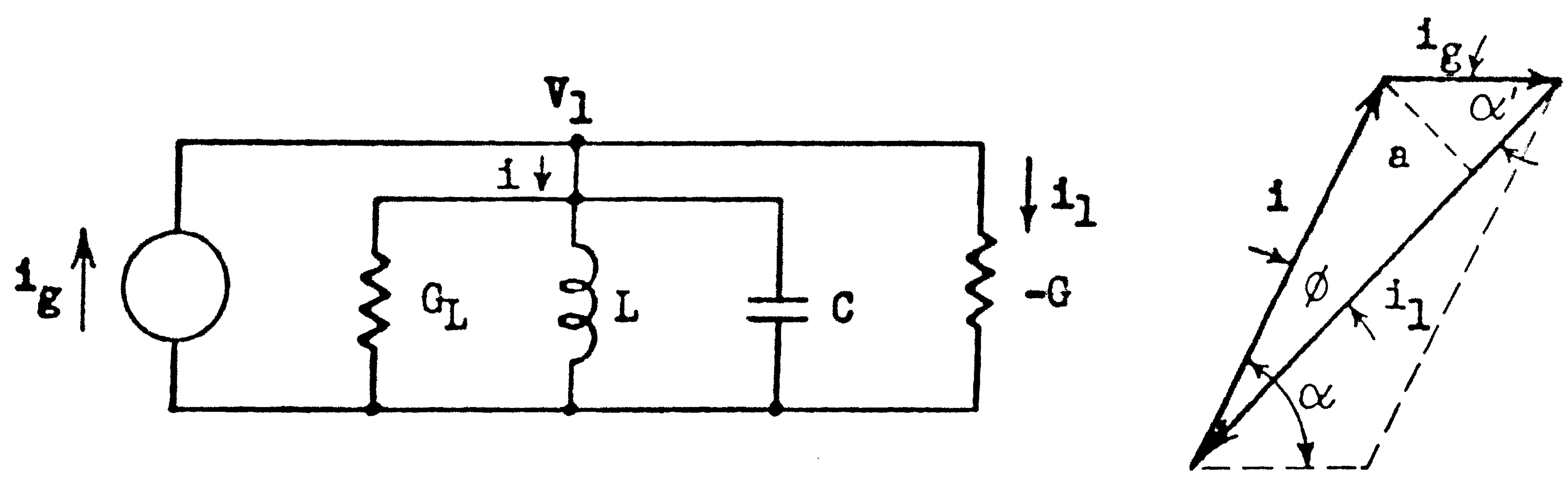


FIGURE 3c. SIMPLIFIED EQUIVALENT CIRCUIT

i_g	=	Locking or Input Signal Current
G_L	=	Transformed Load Impedance
$-G$	=	Linearized Equivalent of $I_1(V_1)$
C	=	Diode and Circuit Capacitance
L	=	Tuning Inductance
$I_1(V_1)$	=	Nonlinear Current-Voltage Characteristic of Diode at Fundamental Frequency of Operation

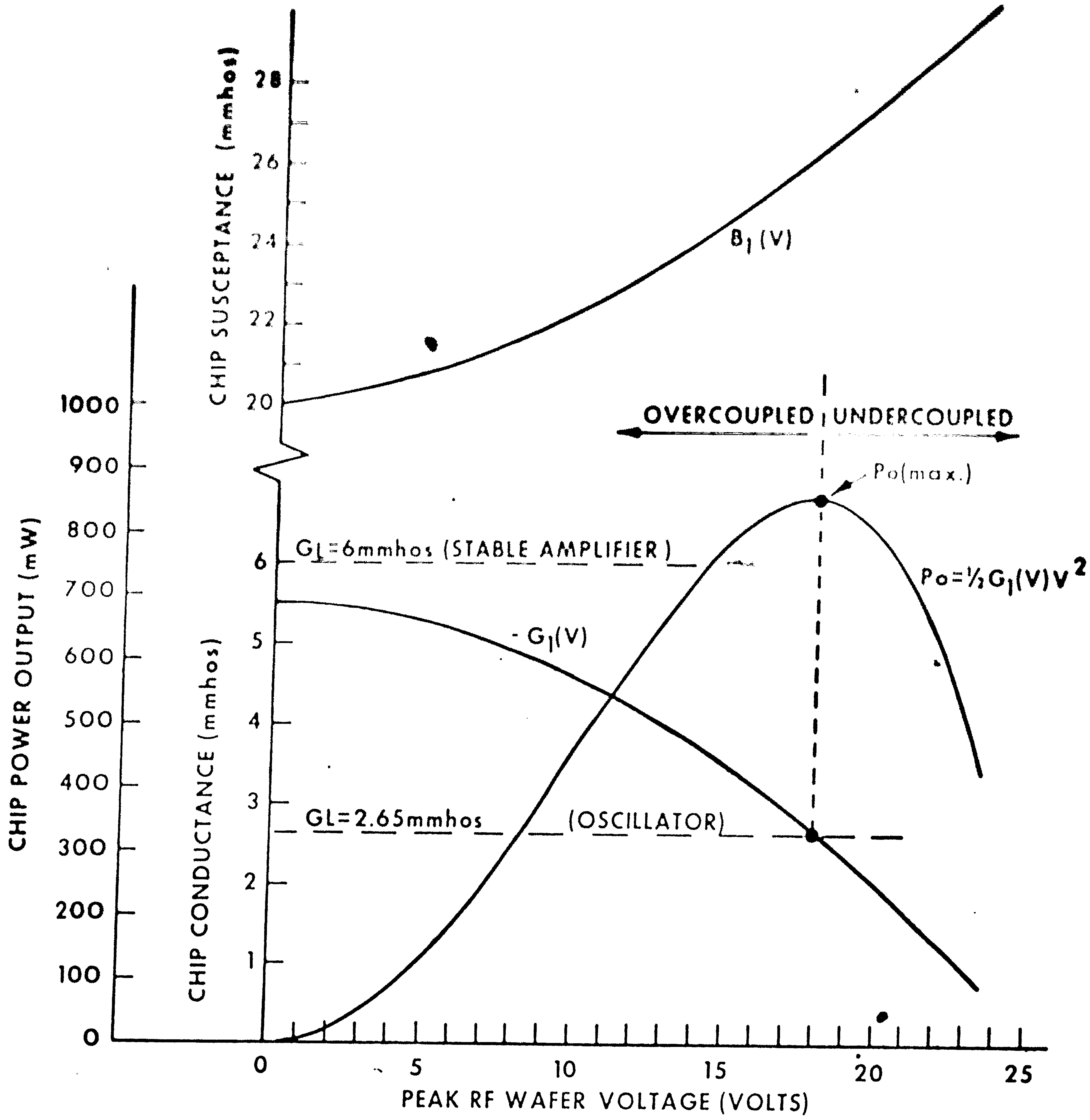


FIGURE 4. ADMITTANCE[10] AND POWER OUTPUT vs RF WAFER VOLTAGE FOR A GERMANIUM IMPATT DIODE AT A PARTICULAR BIAS CURRENT AND OPERATING FREQUENCY

FOR $G_L + G_1(V) > 0$, THE DIODE WILL OPERATE AS A STABLE AMPLIFIER.

FOR $G_L + G_1(V) < 0$, THE DIODE WILL OPERATE AS AN OSCILLATOR.

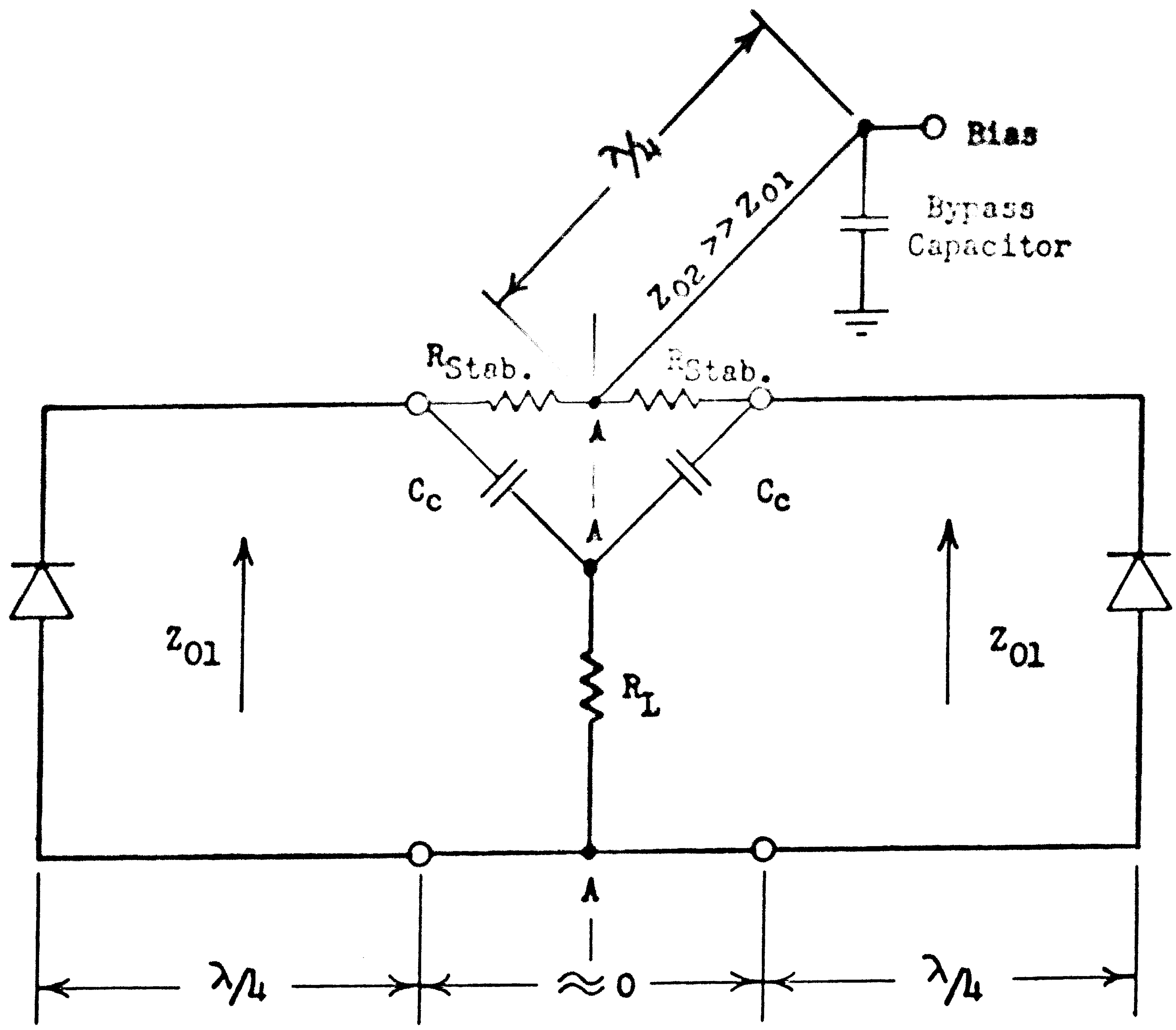


FIGURE 5. A SCHEMATIC DIAGRAM OF TWO SECTIONS OF C.T.RUCKER'S POWER COMBINER [16]

Additional Sections may be added at Points "A"

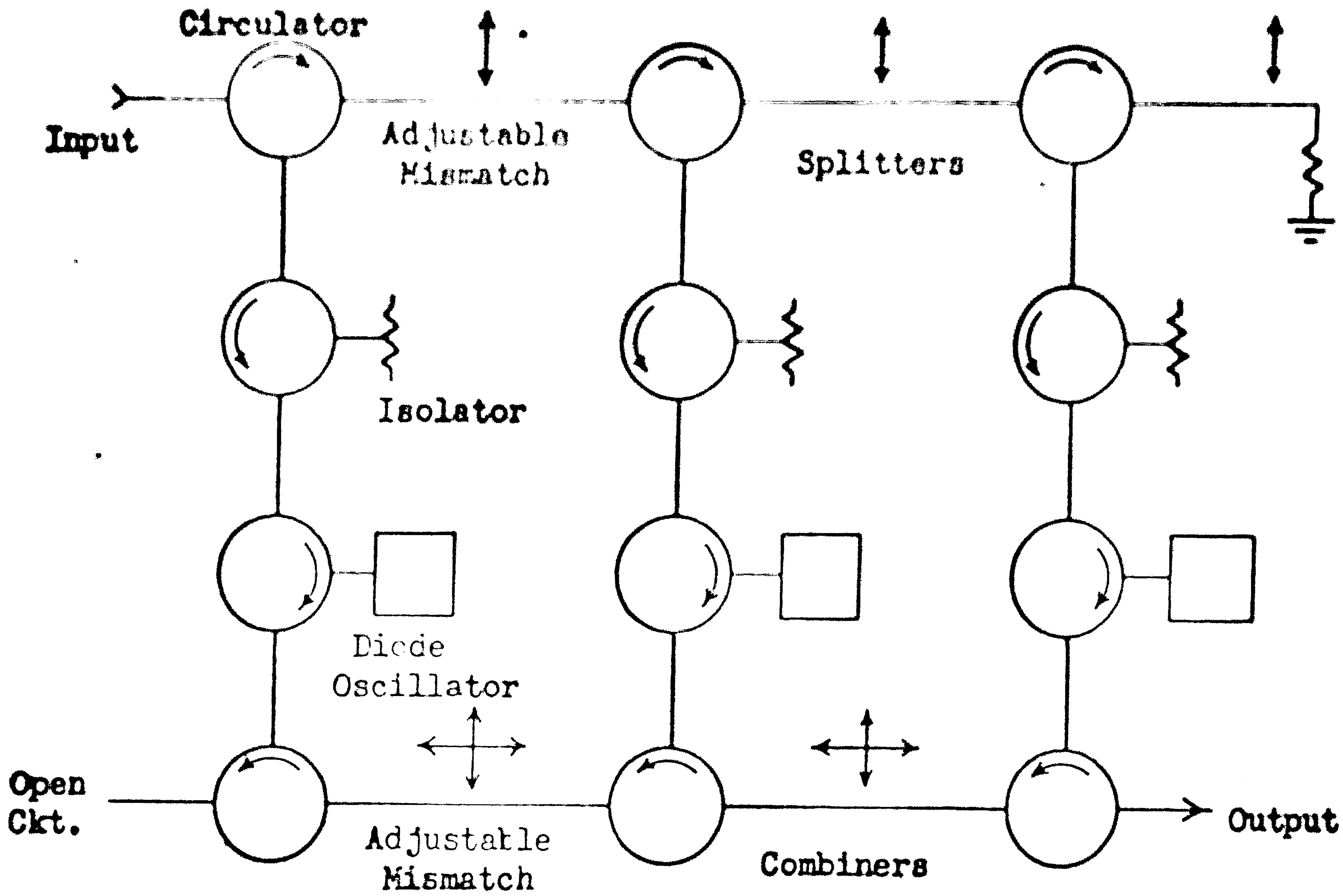


FIGURE 6a. A THREE SECTION CIRCULATOR POWER COMBINER [19]

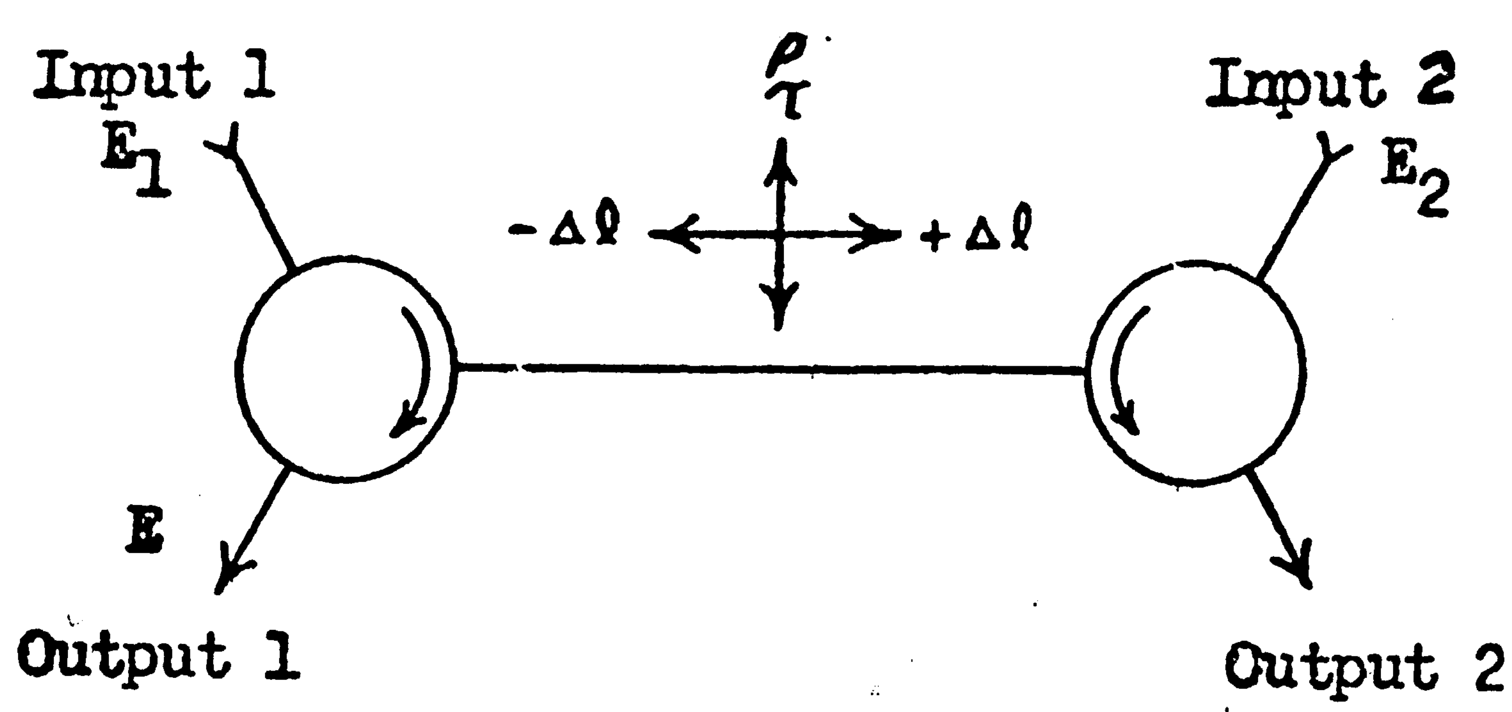


FIGURE 6b. A SINGLE COMBINER-SPLITTER SECTION

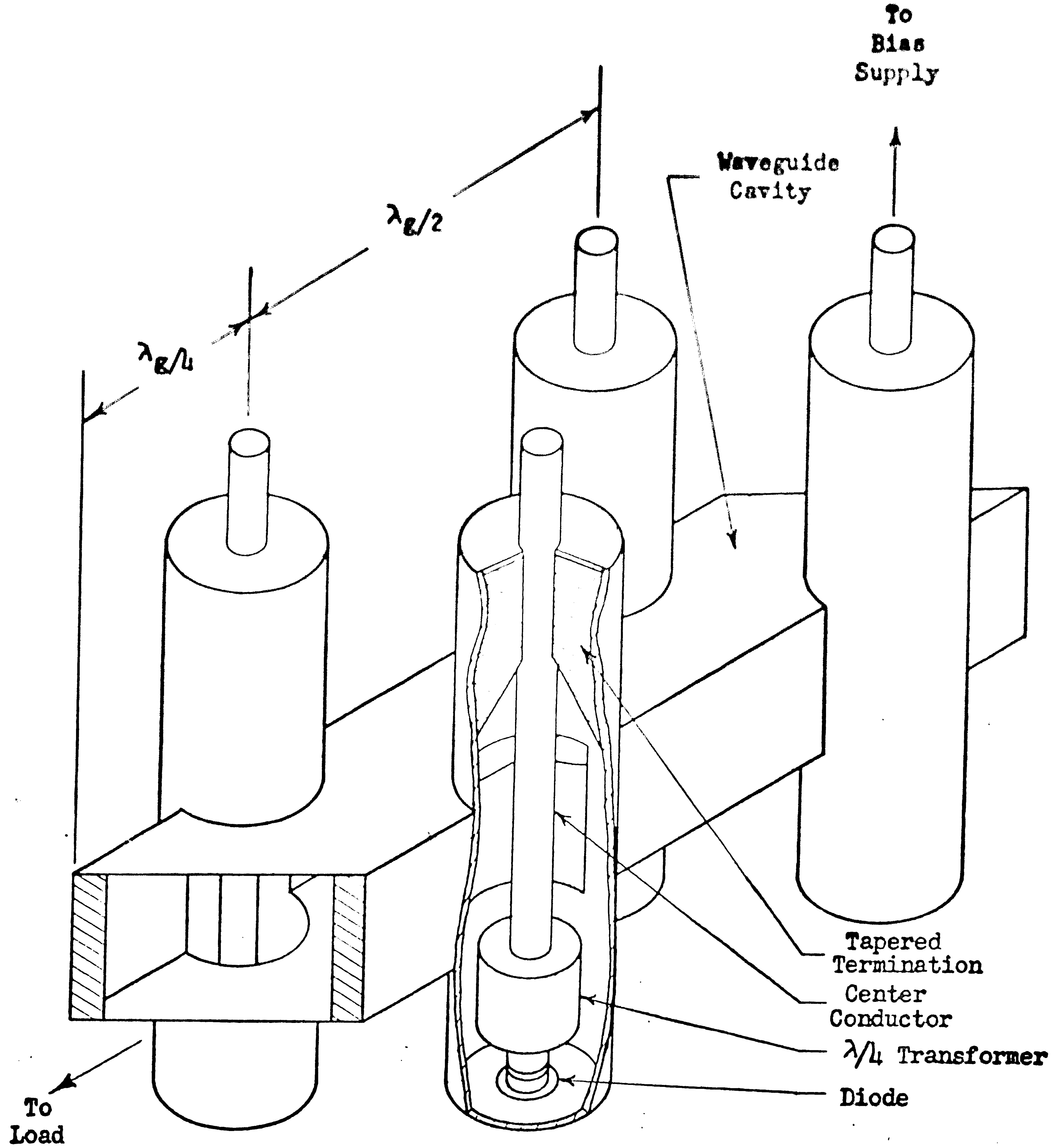


FIGURE 7. SINGLE CAVITY WAVEGUIDE COMBINER [21]

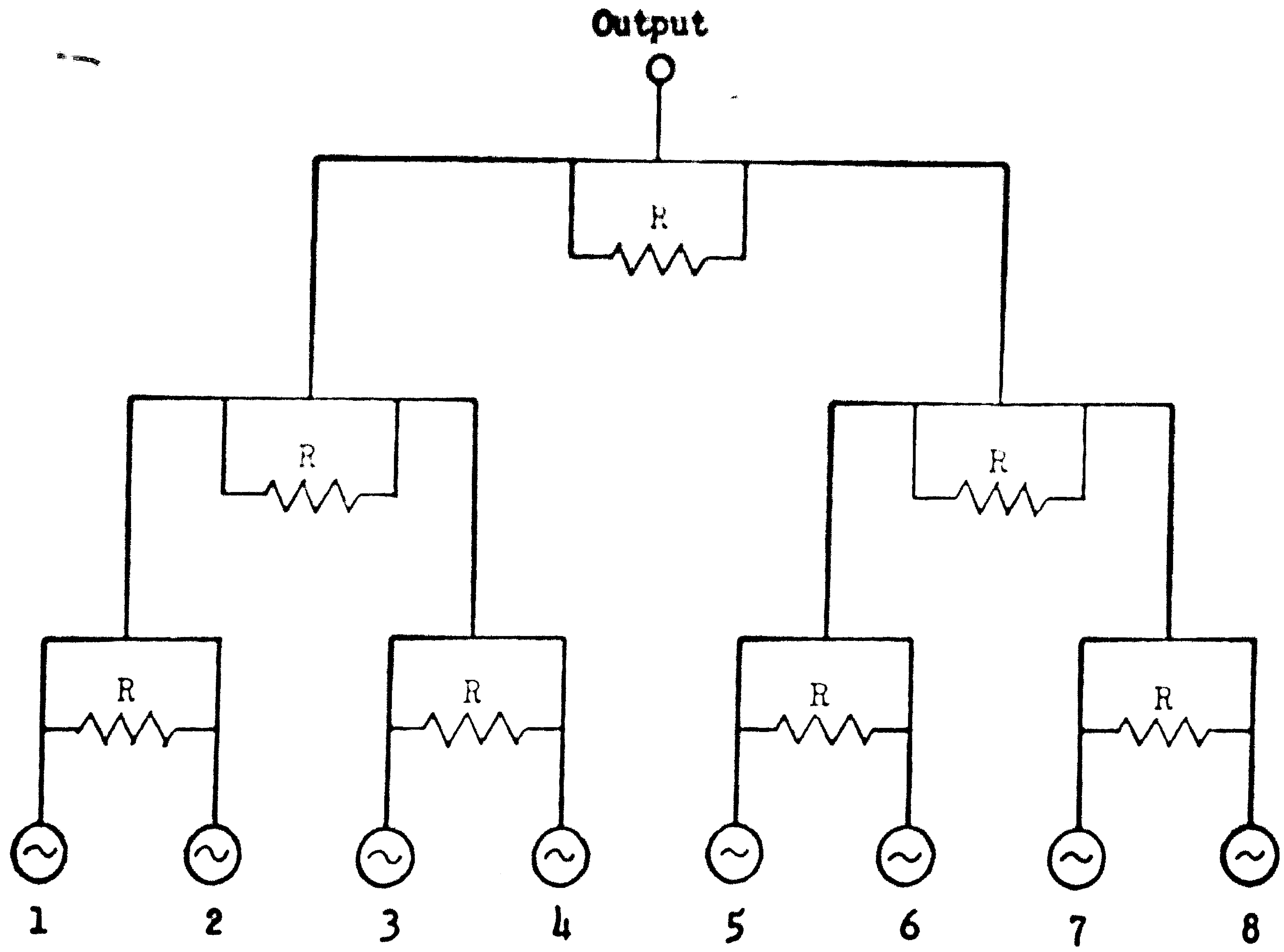


FIGURE 8a. RESISTIVE HYBRID COMBINER [23]

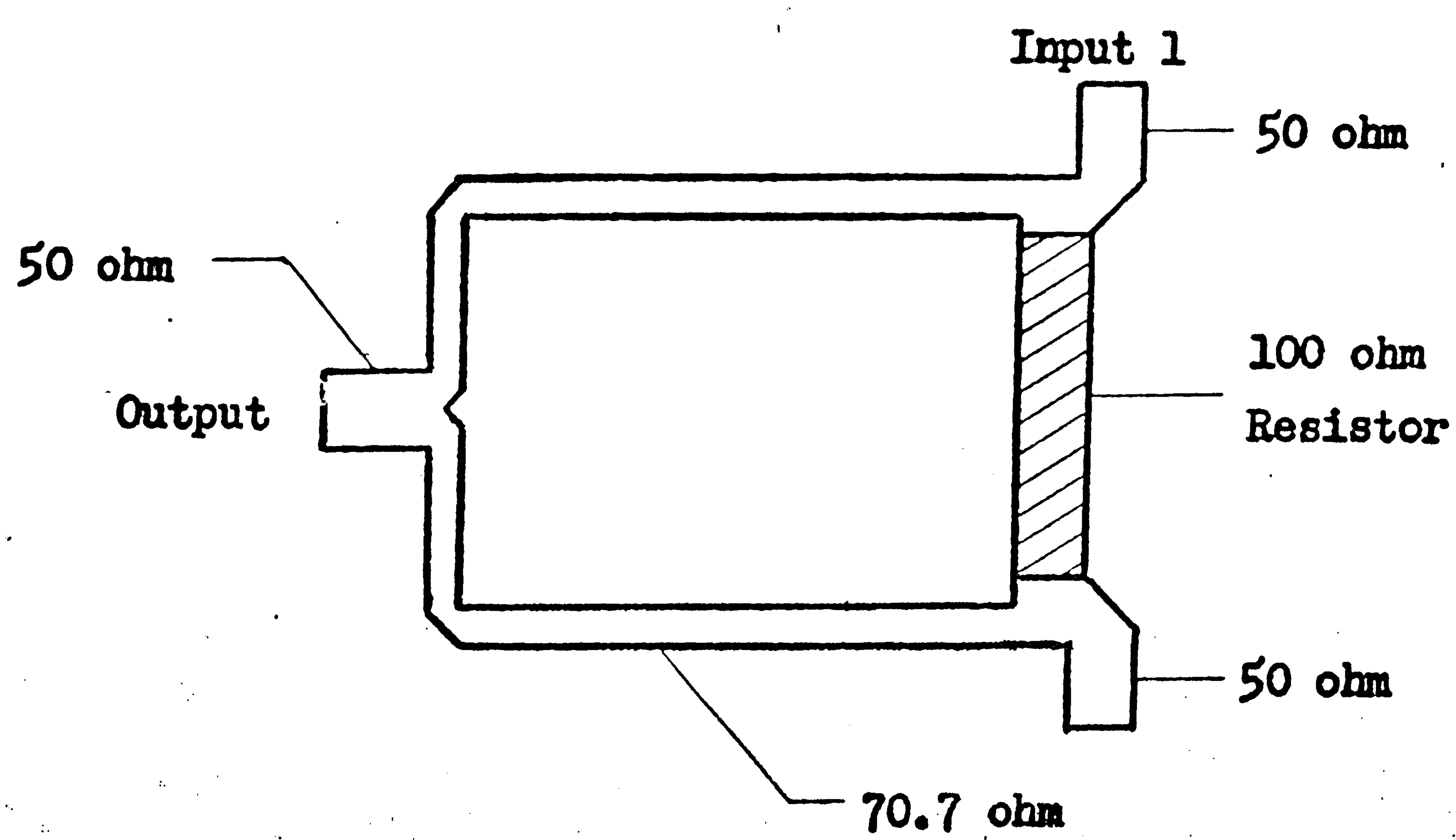


FIGURE 8b. SINGLE HYBRID CIRCUIT [24]

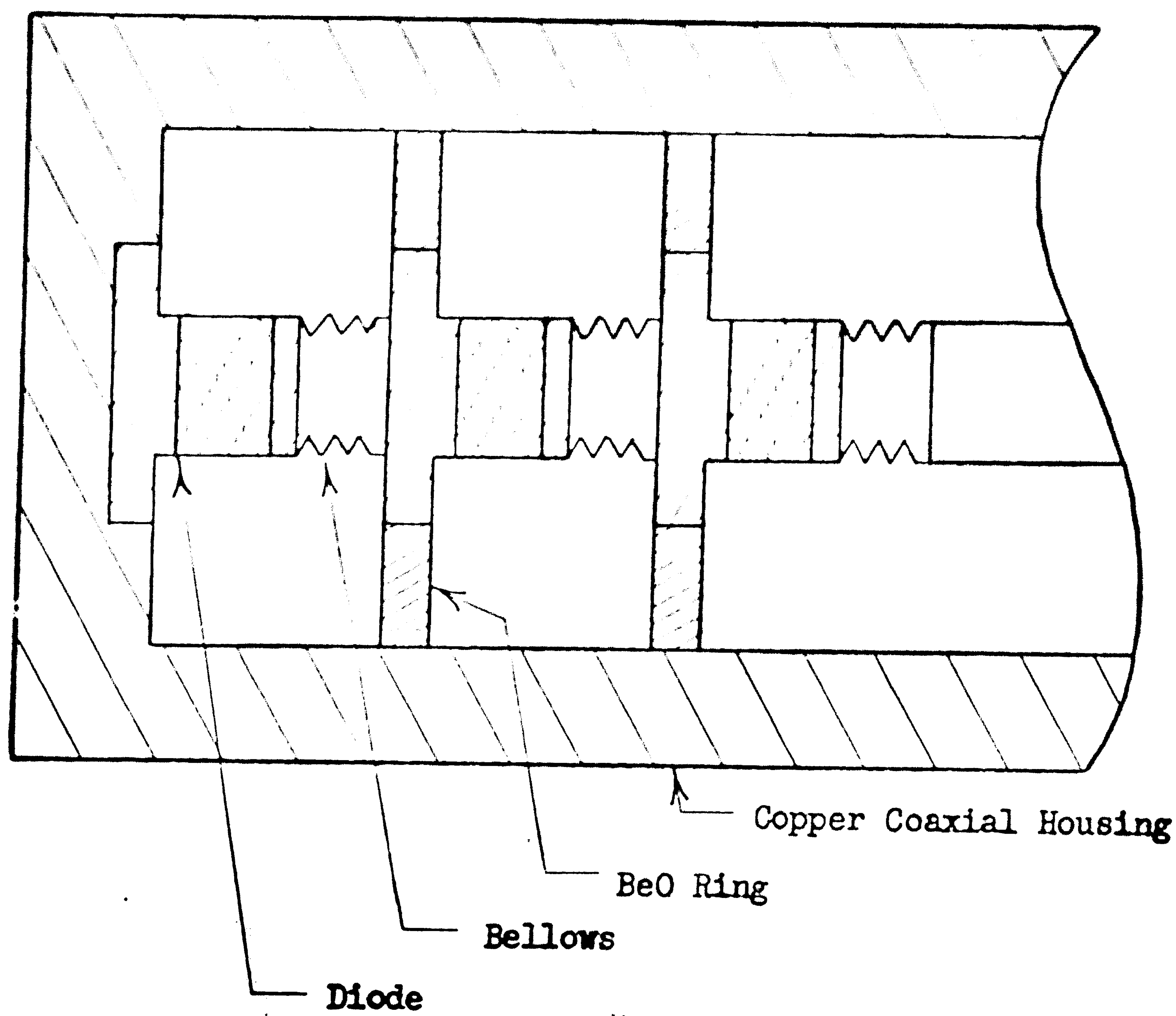


FIGURE 9. A SERIES COMBINATION OF PACKAGED DIODES [25]

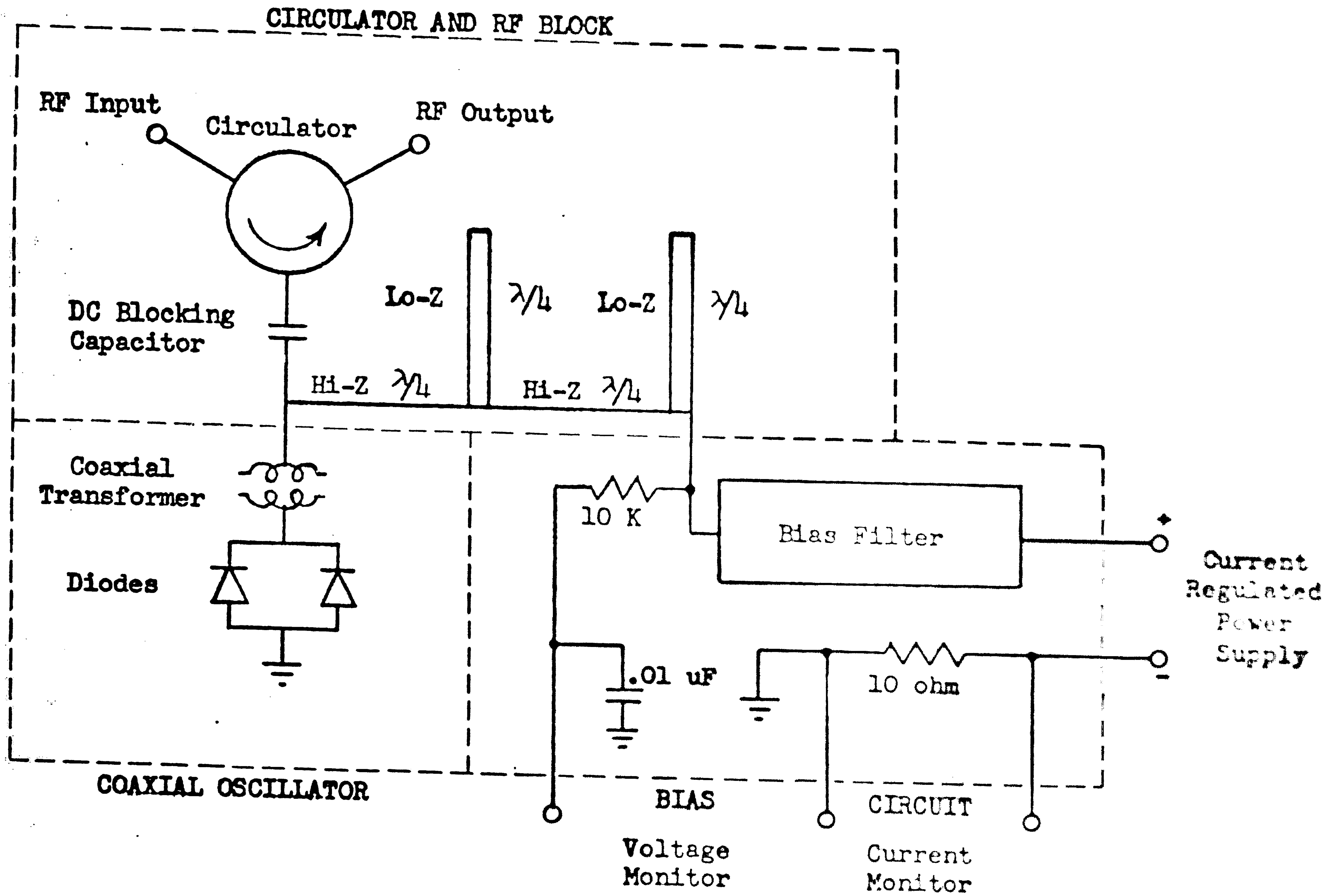


FIGURE 10. SCHEMATIC DIAGRAM OF PARALLEL DIODE OSCILLATOR

7mm COAXIAL IMPATT OSCILLATOR

USING IMPATT PAIR

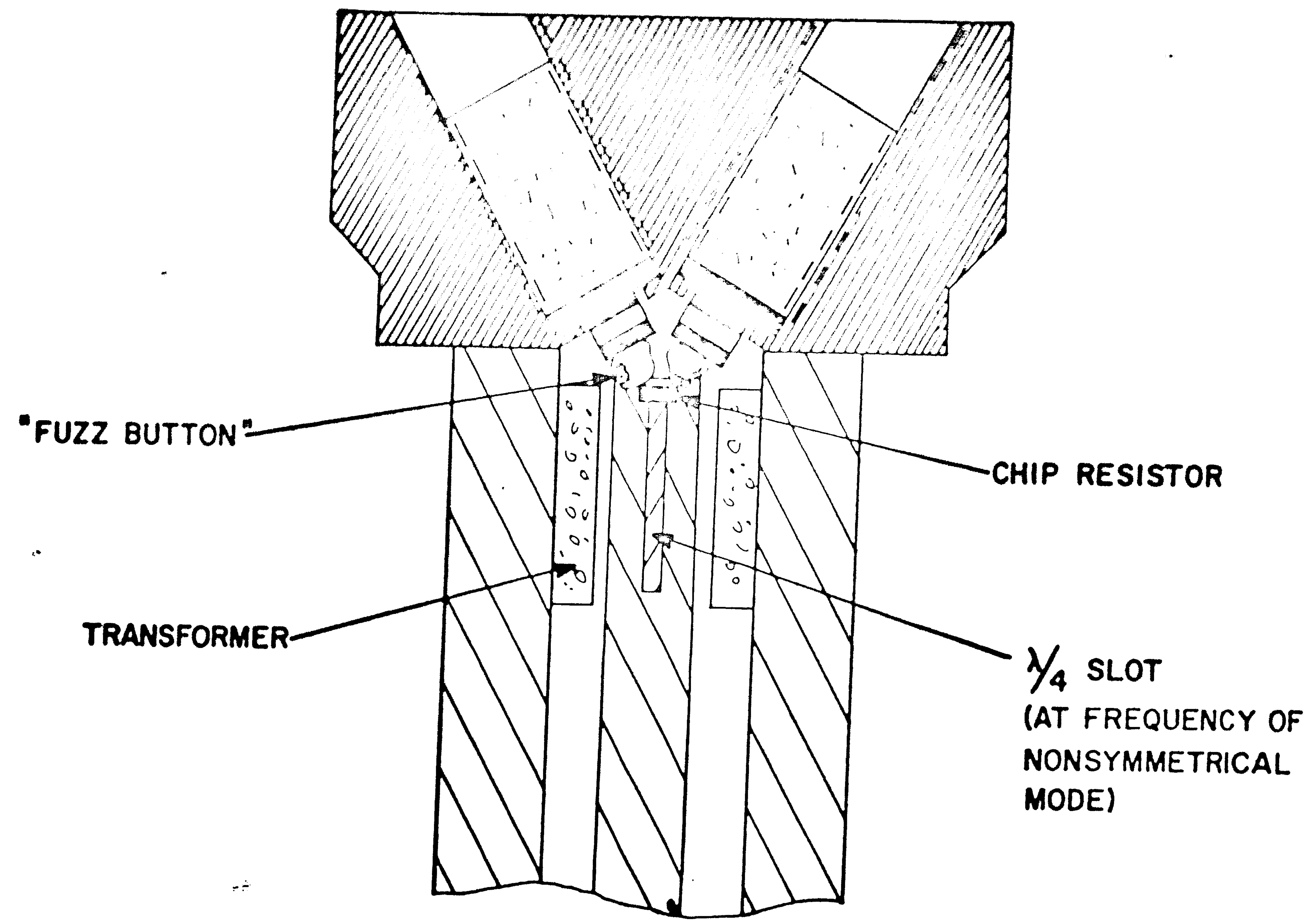


FIGURE 11

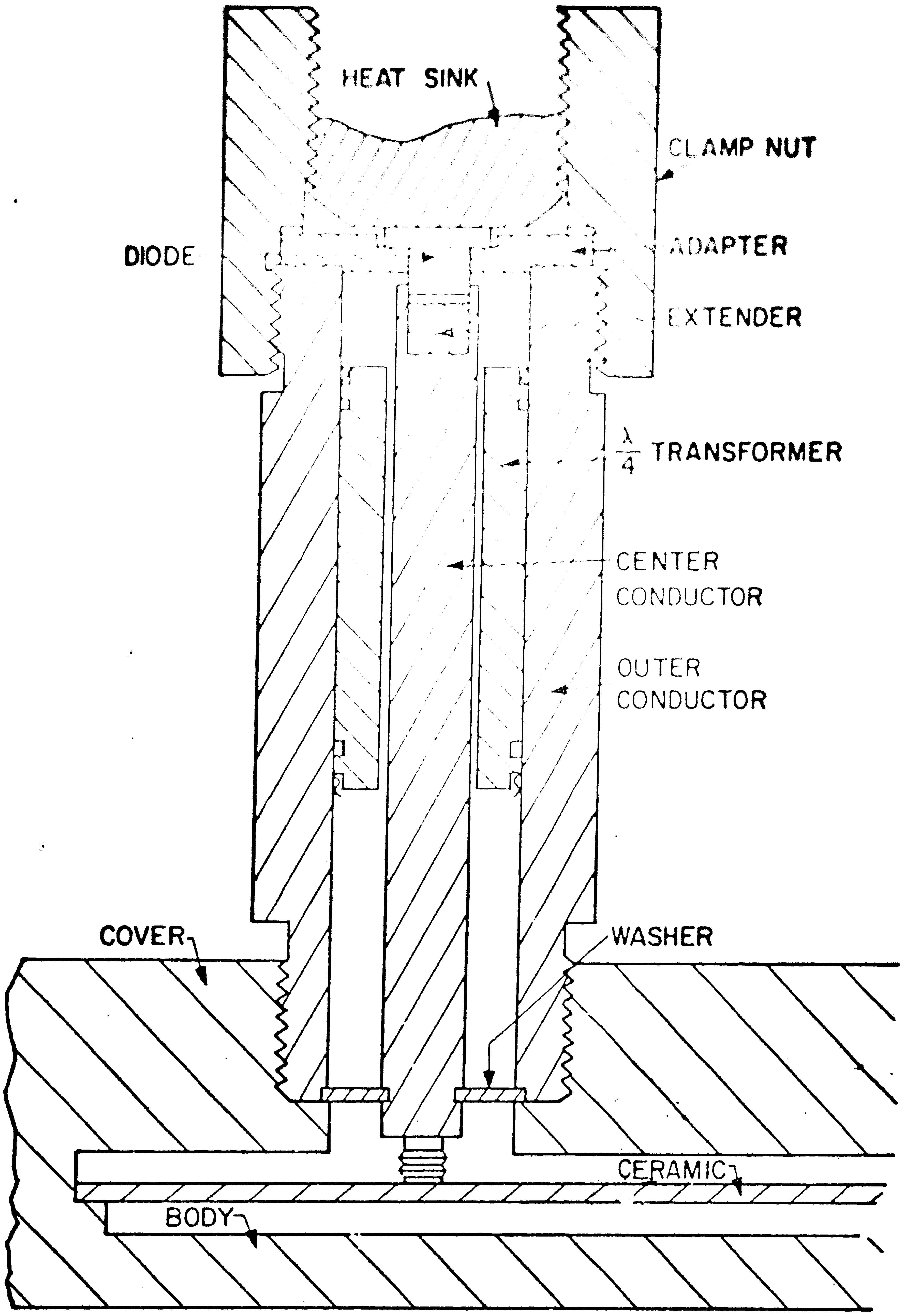


FIGURE 12. SINGLE DIODE COAXIAL HOUSING

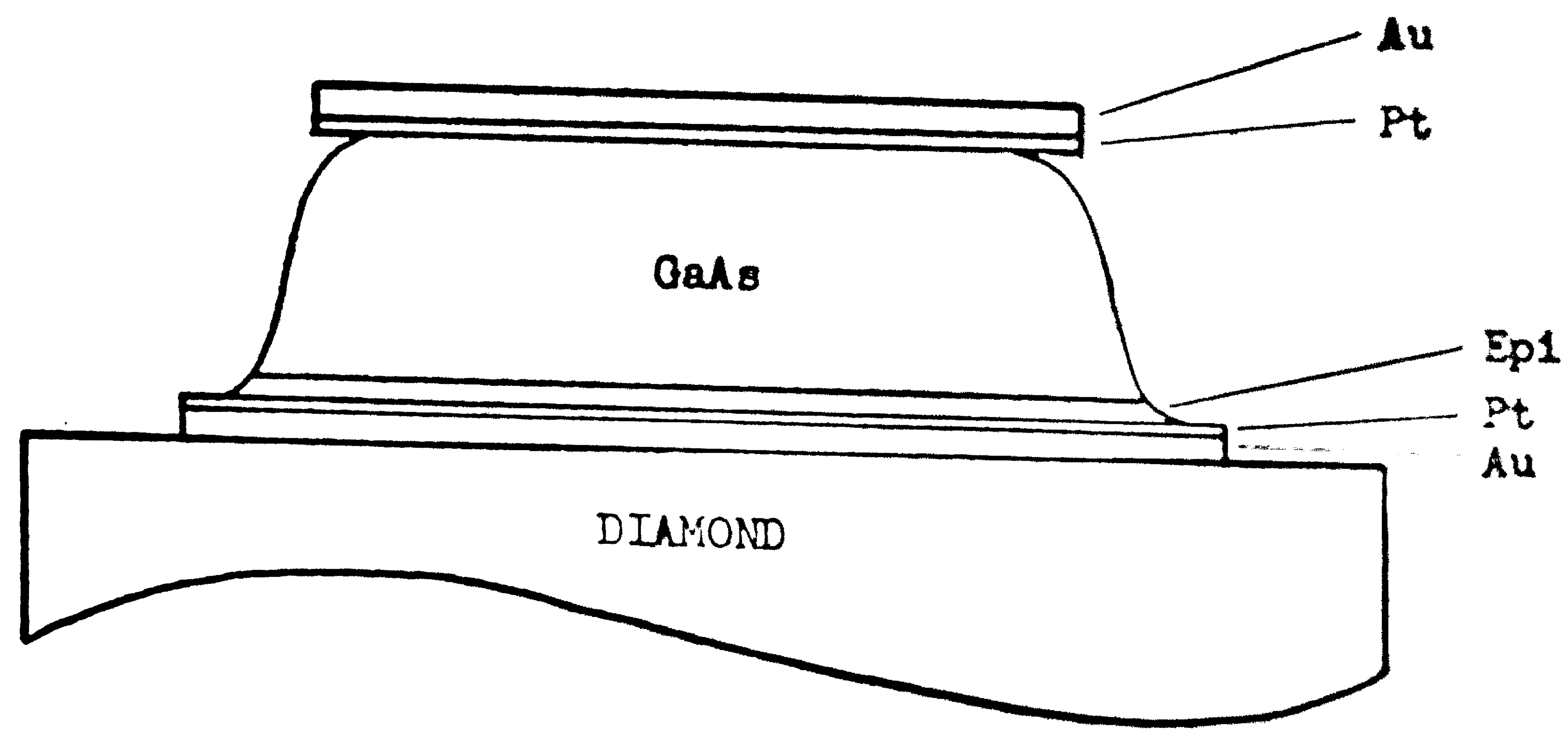
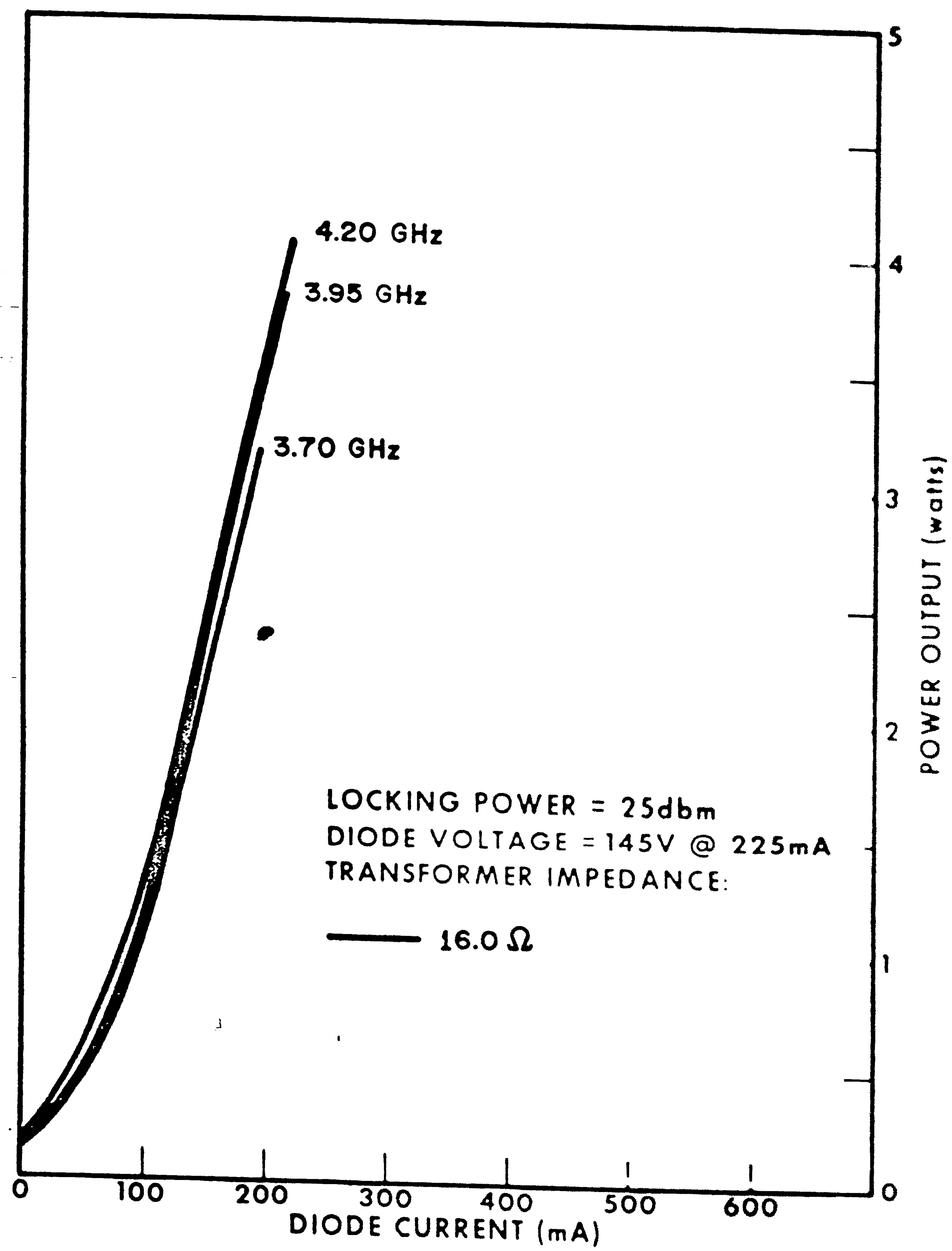


FIGURE 13. GALLIUM ARSENIDE SCHOTTKY BARRIER IMPATT DIODE

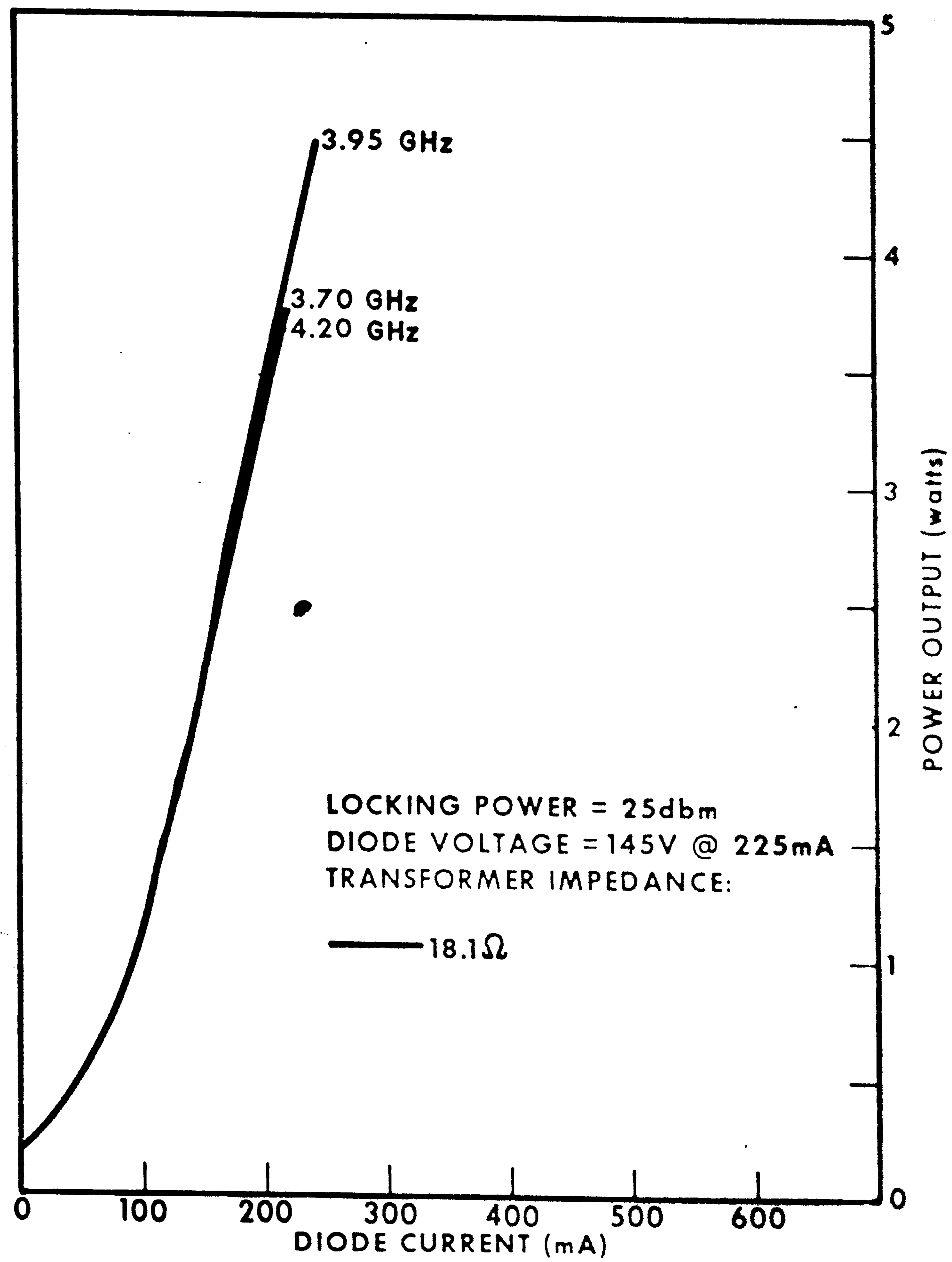
The active region of the diode is entirely contained within the Epitaxial Layer.



DIODE POWER OUTPUT vs. CURRENT

DIODE: # 1

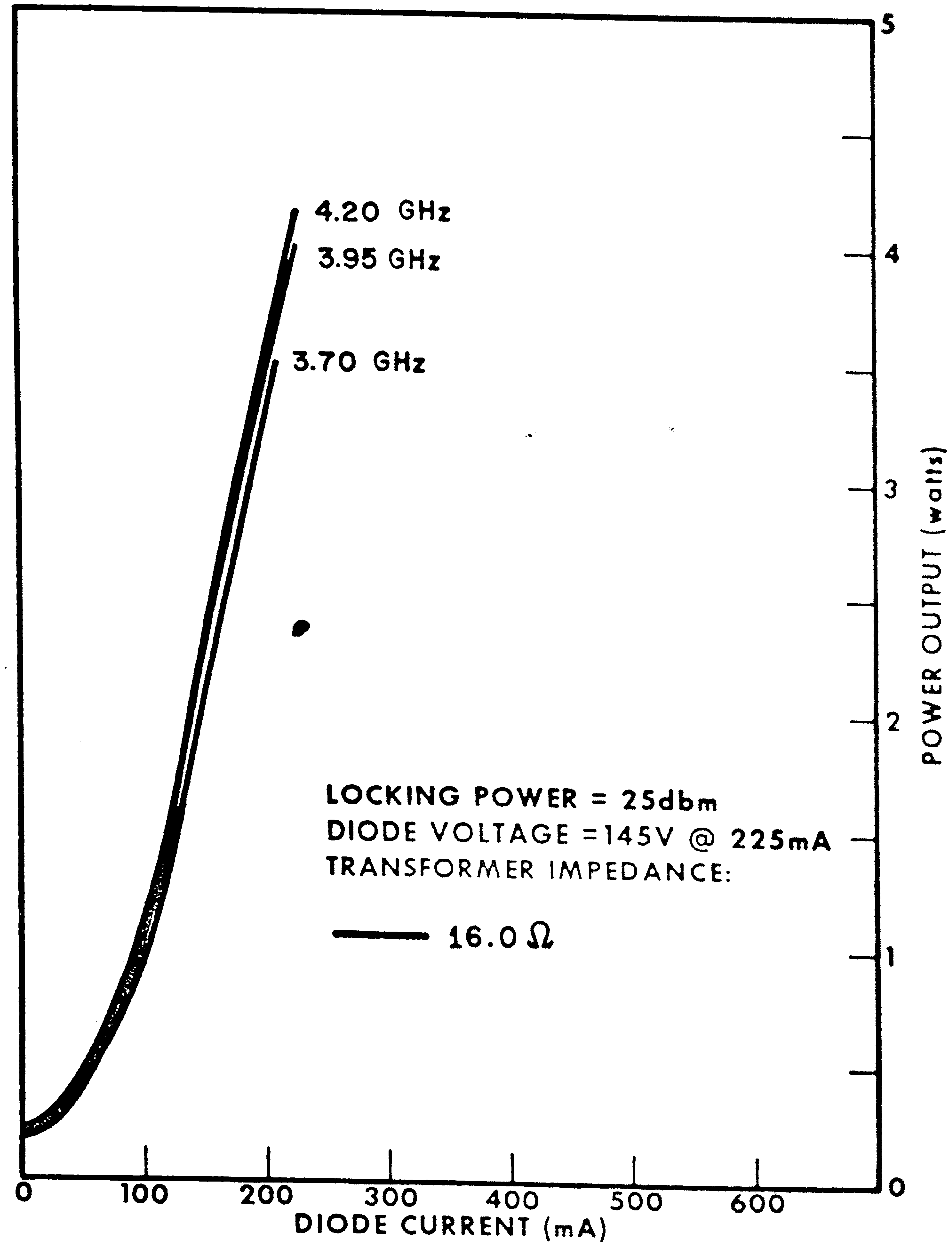
FIGURE 14



DIODE POWER OUTPUT vs. CURRENT

DIODE: # 1

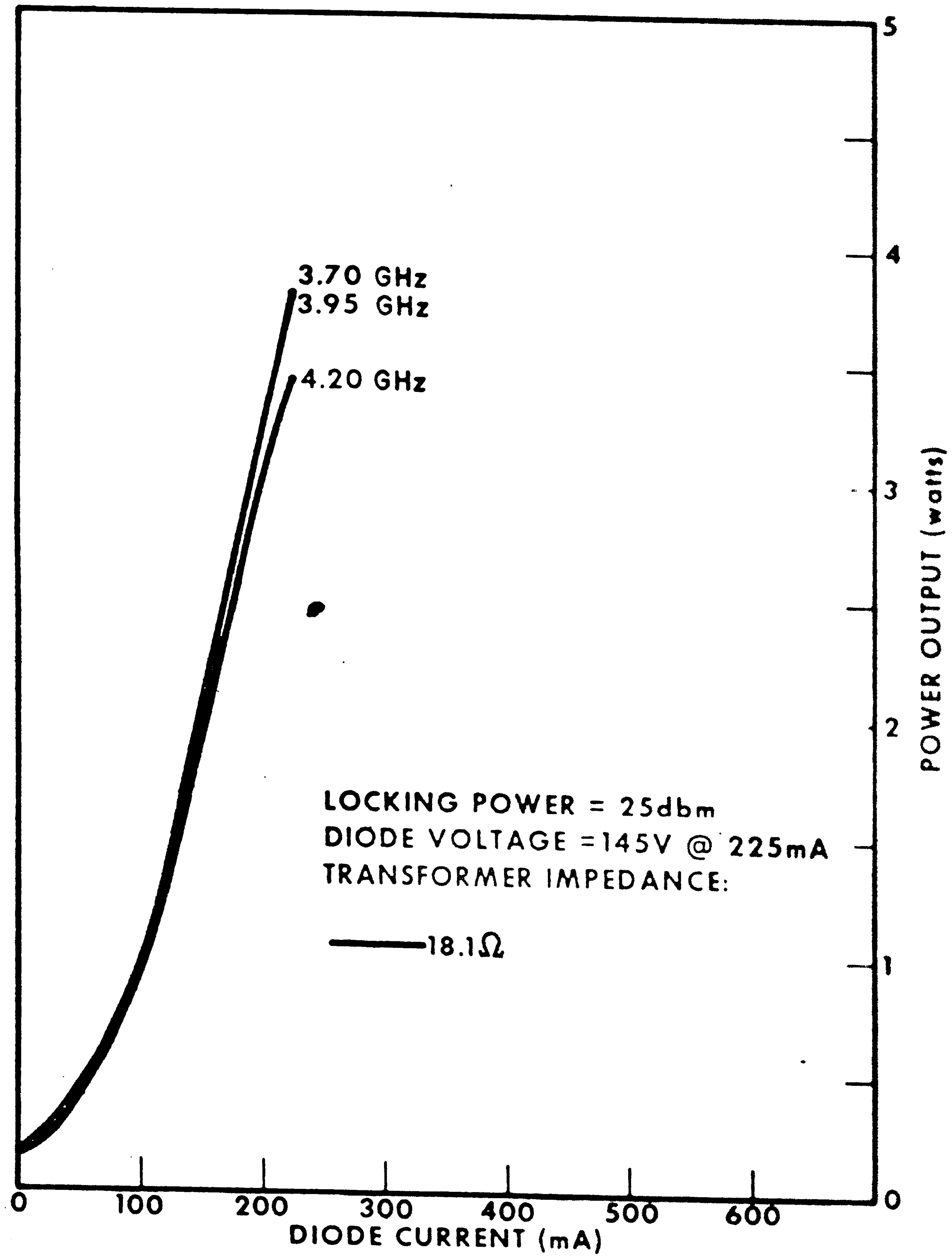
FIGURE 15



DIODE POWER vs. CURRENT

DIODE: # 2

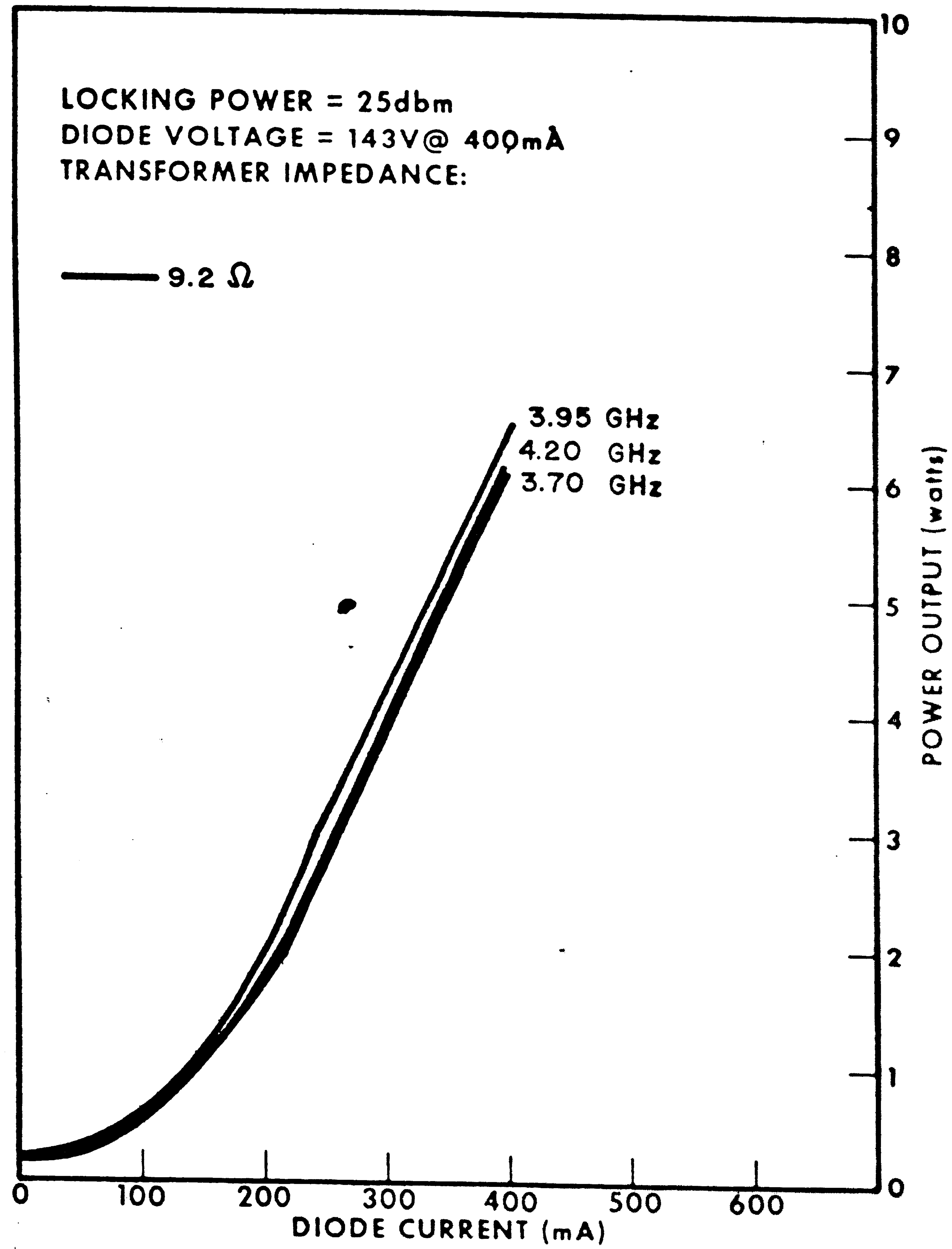
FIGURE 16



DIODE POWER vs. CURRENT

DIODE: # 2

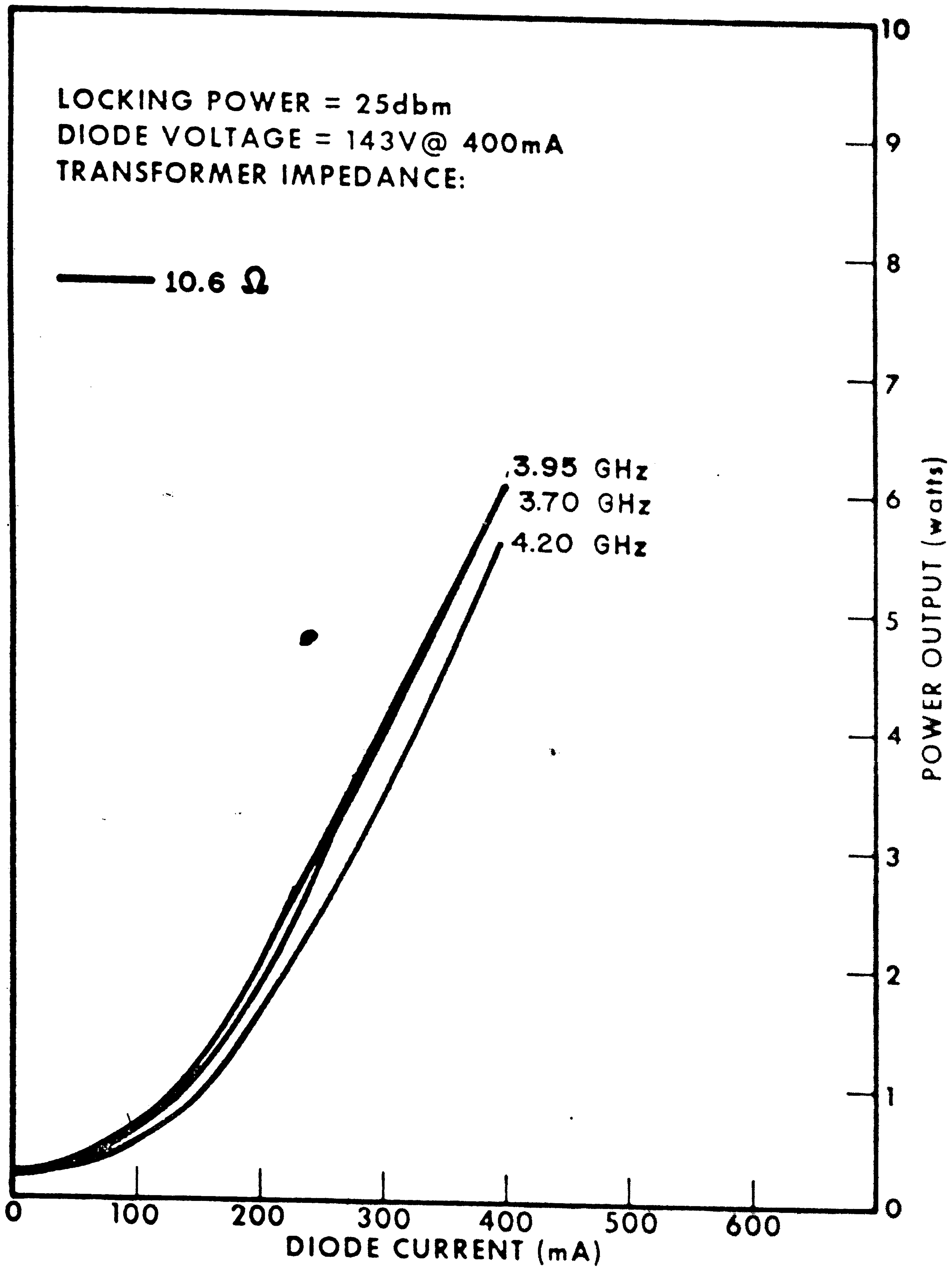
FIGURE 17



DIODE POWER vs. CURRENT FOR DIODE PAIR

DIODES: 1 AND 2

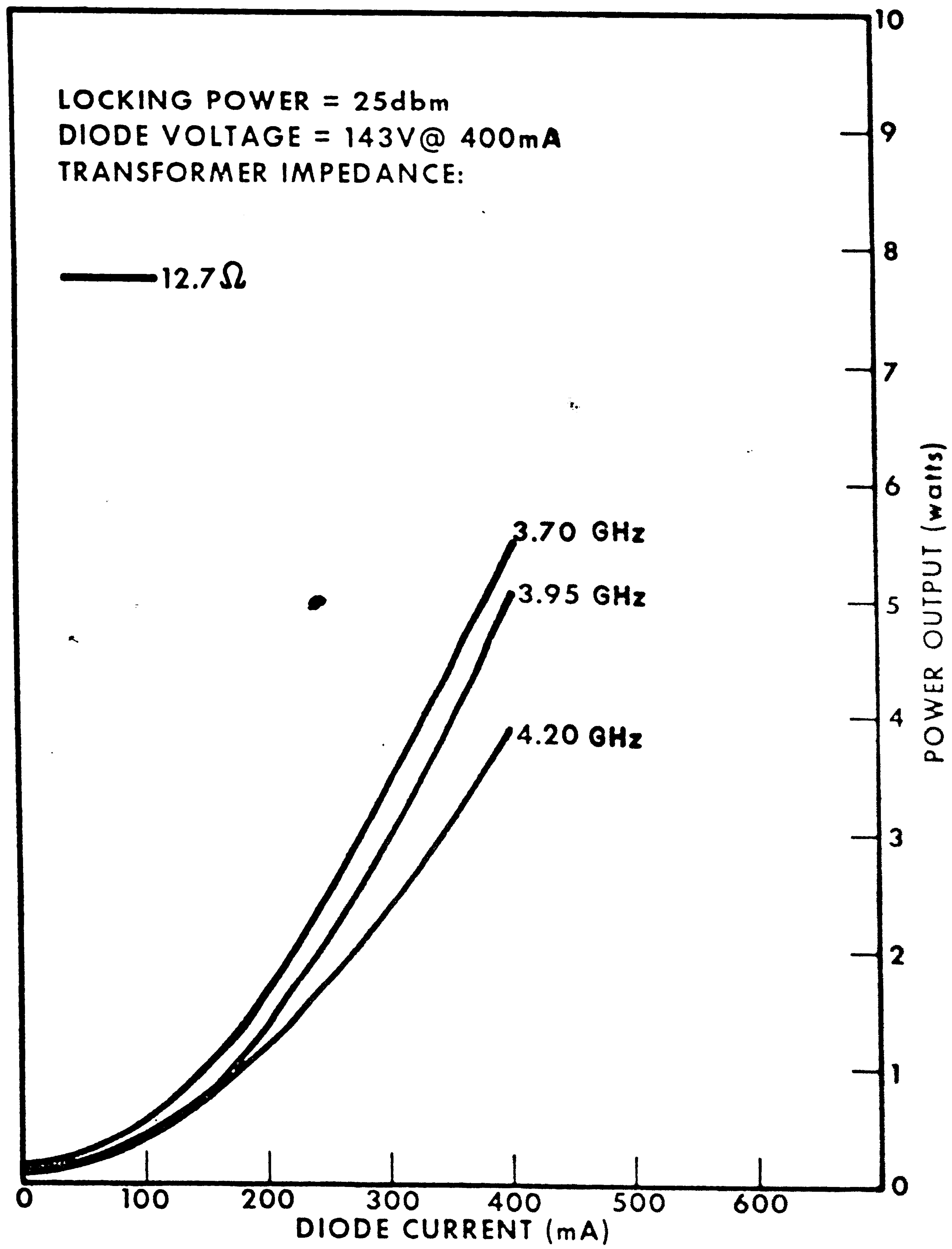
FIGURE 18



DIODE POWER vs. CURRENT FOR DIODE PAIR

DIODES: 1 AND 2

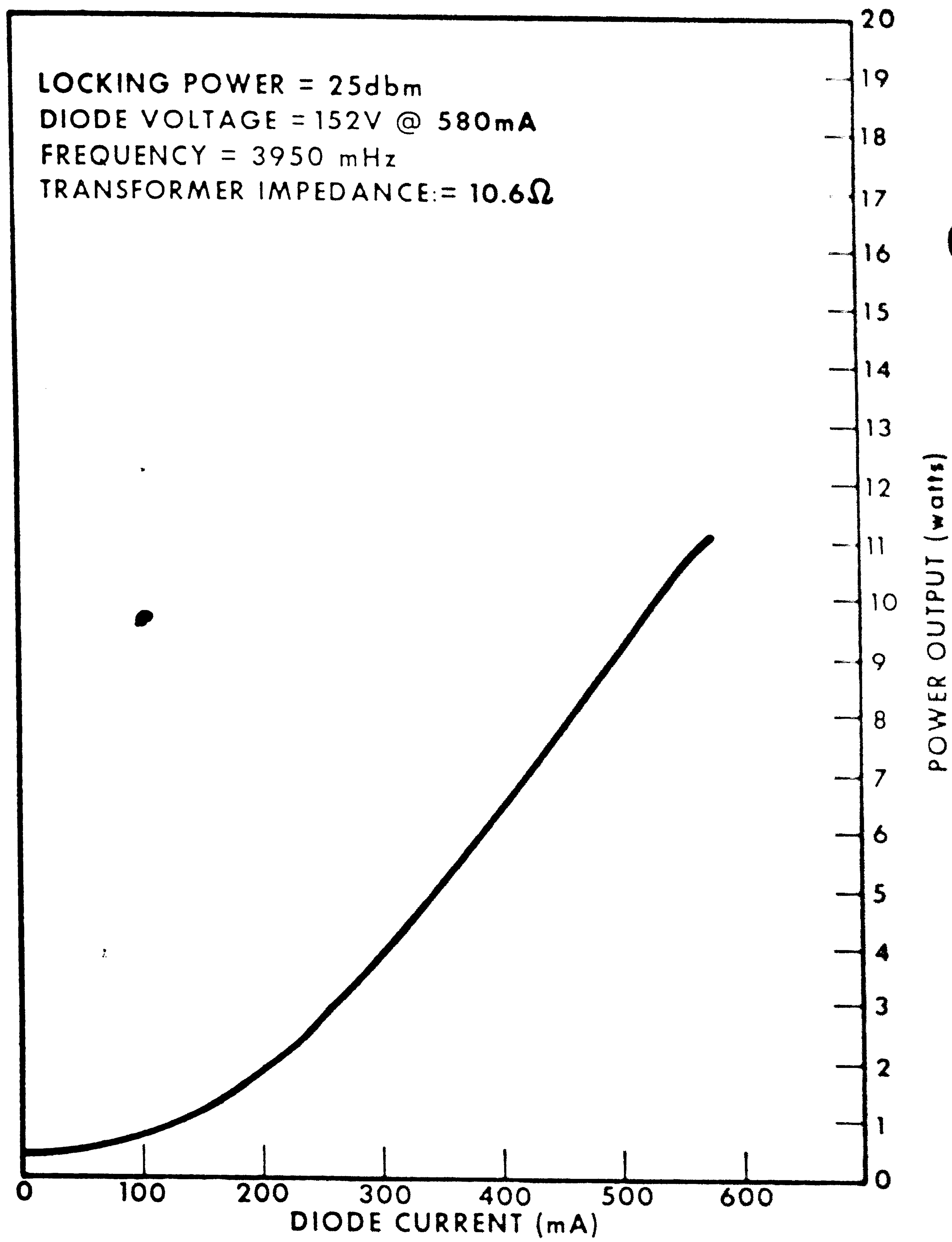
FIGURE 19



DIODE POWER vs. CURRENT FOR DIODE PAIR

DIODES: 1 AND 2

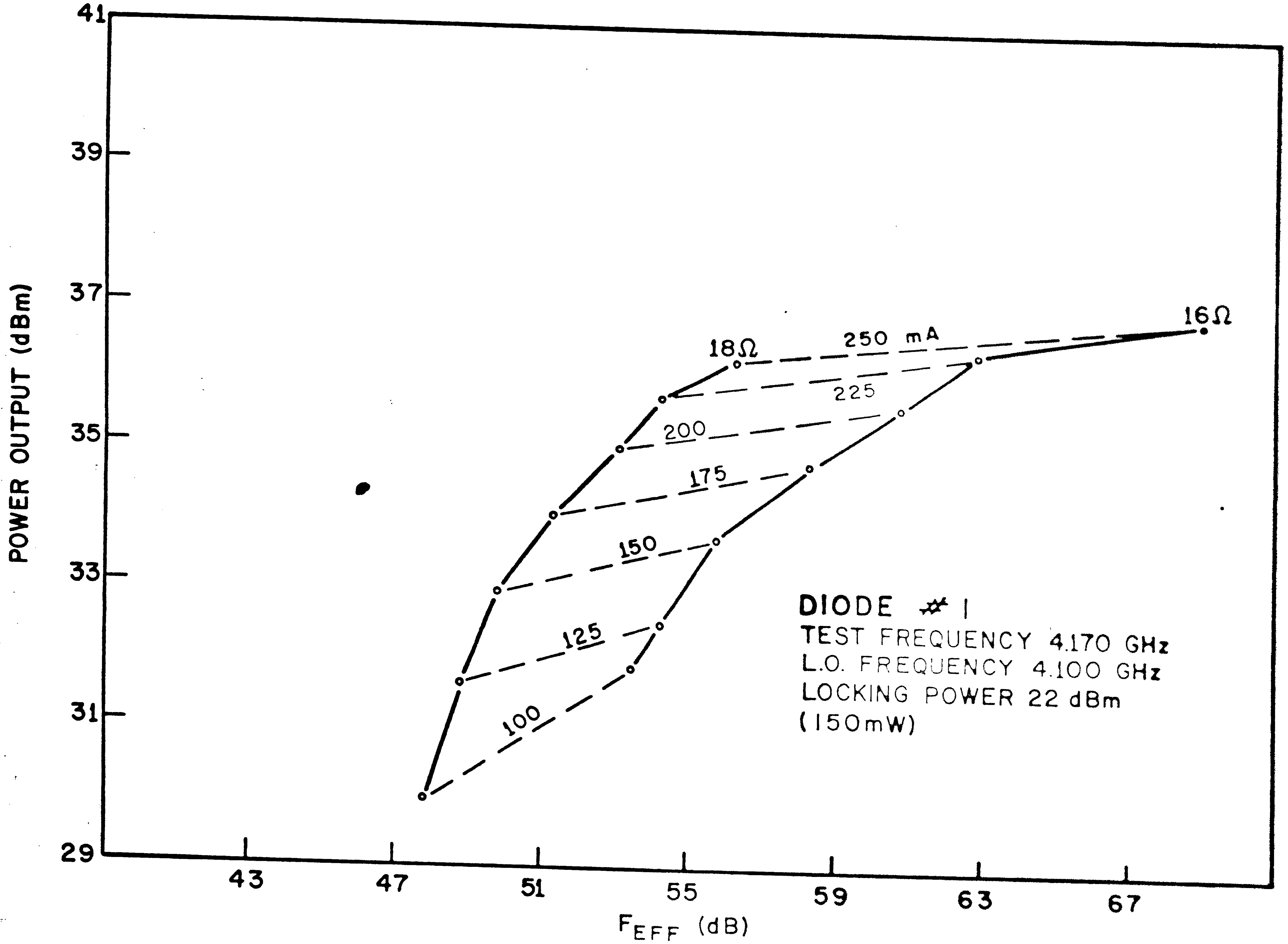
FIGURE 20



DIODE PAIR POWER OUTPUT vs. CURRENT

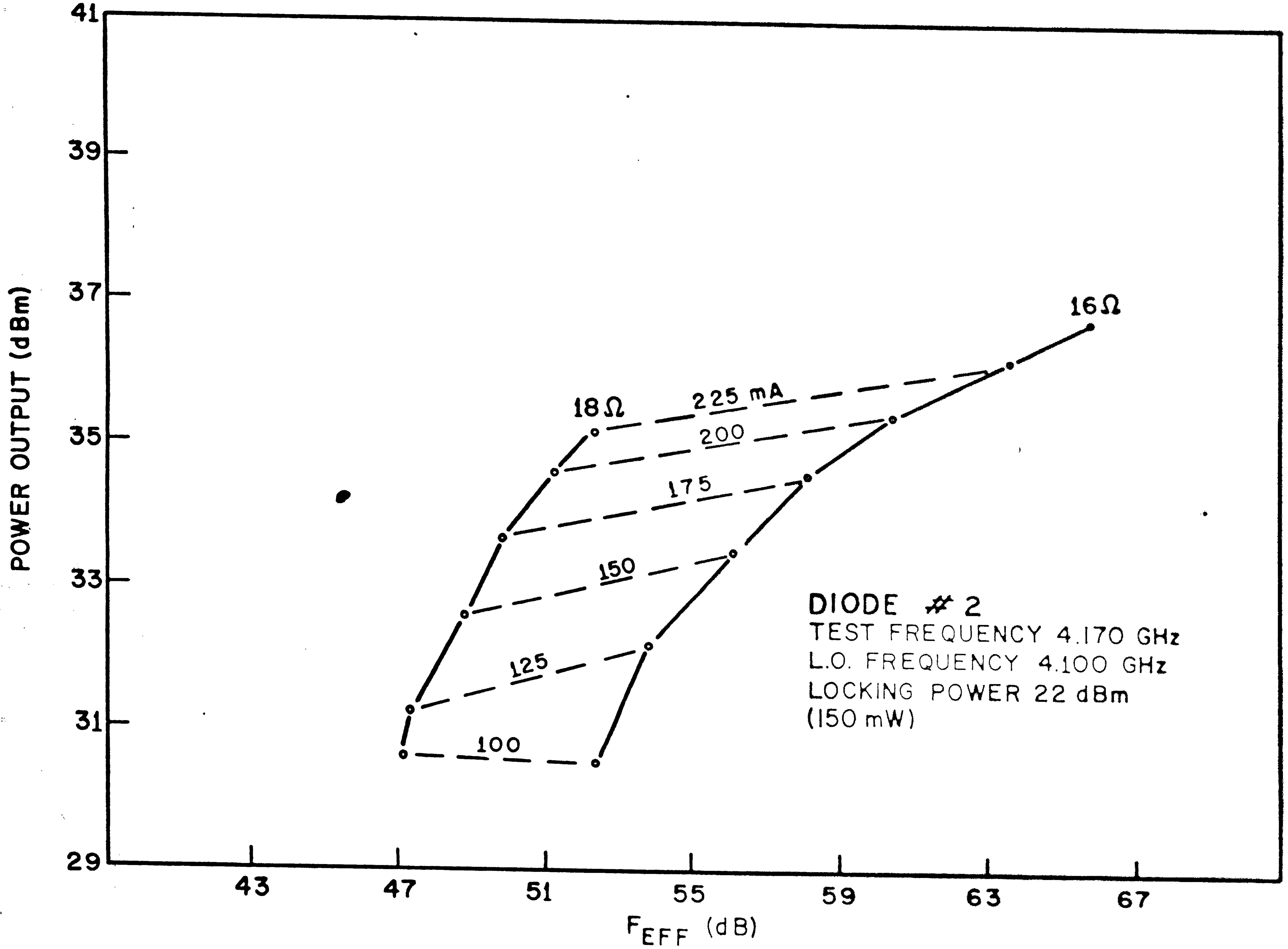
DIODES: 1 AND 2

FIGURE 21



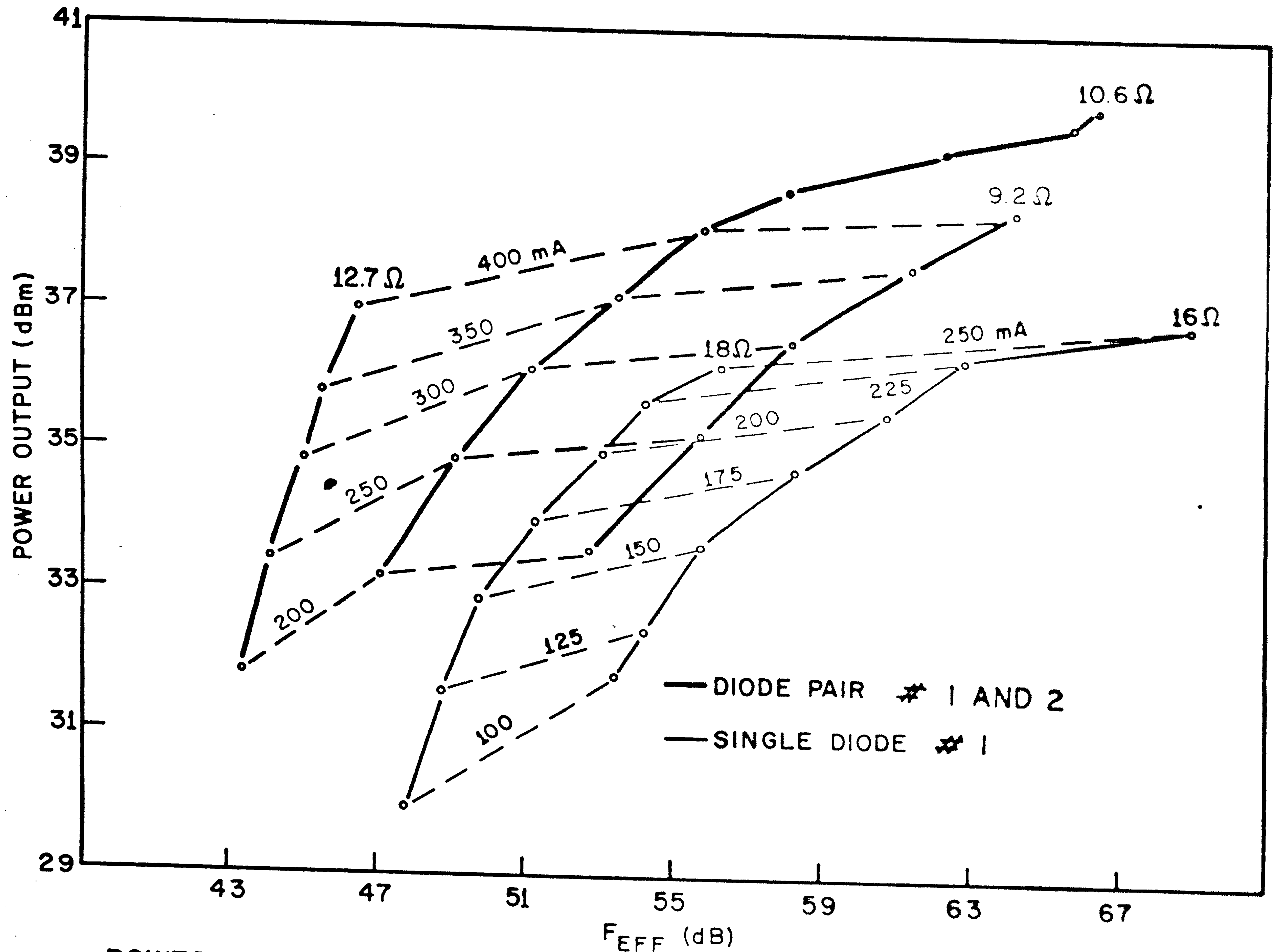
POWER-NOISE PERFORMANCE

FIGURE 22



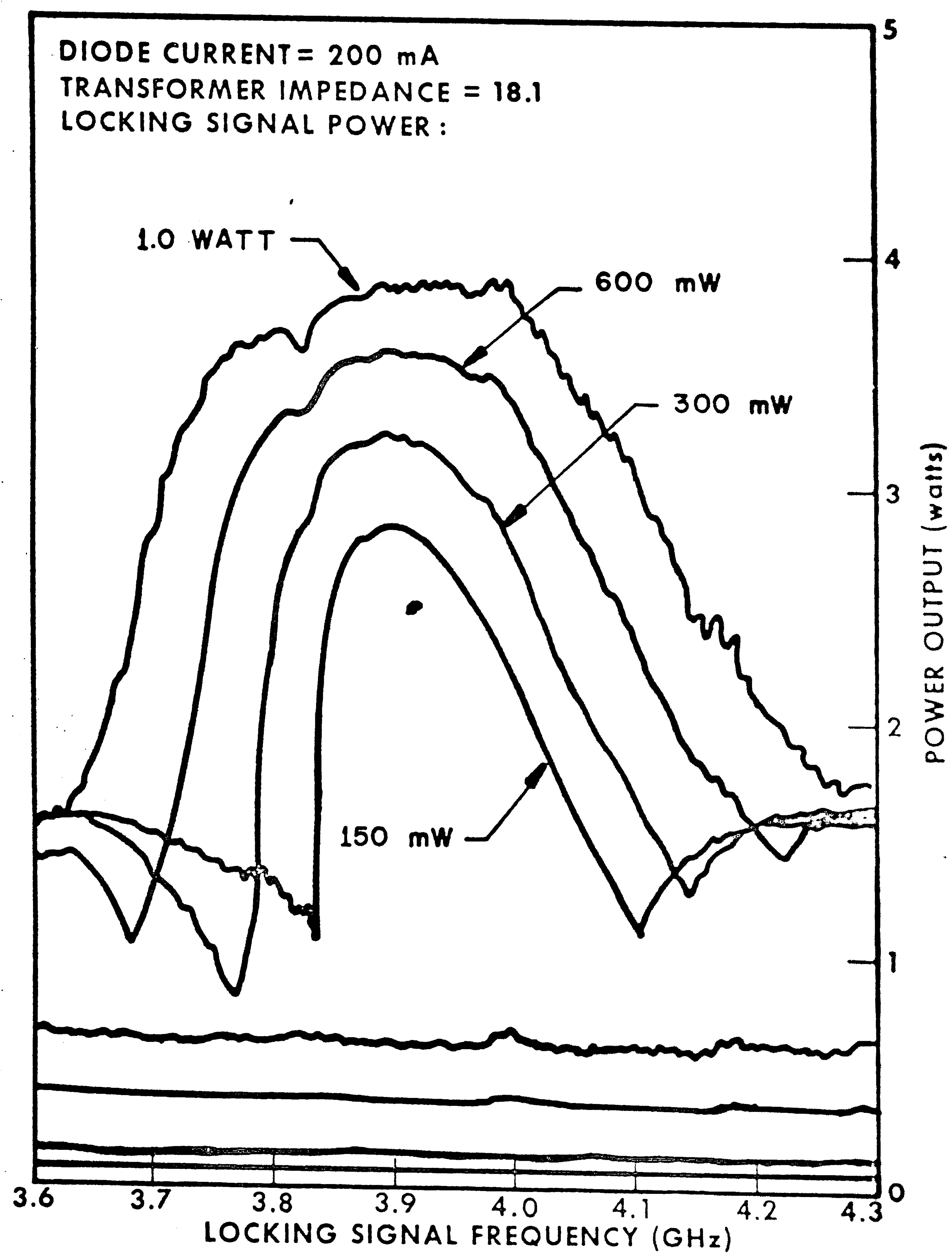
POWER-NOISE PERFORMANCE

FIGURE 23



POWER VS NOISE FOR DIODE PAIR AS COMPARED TO SINGLE DIODE

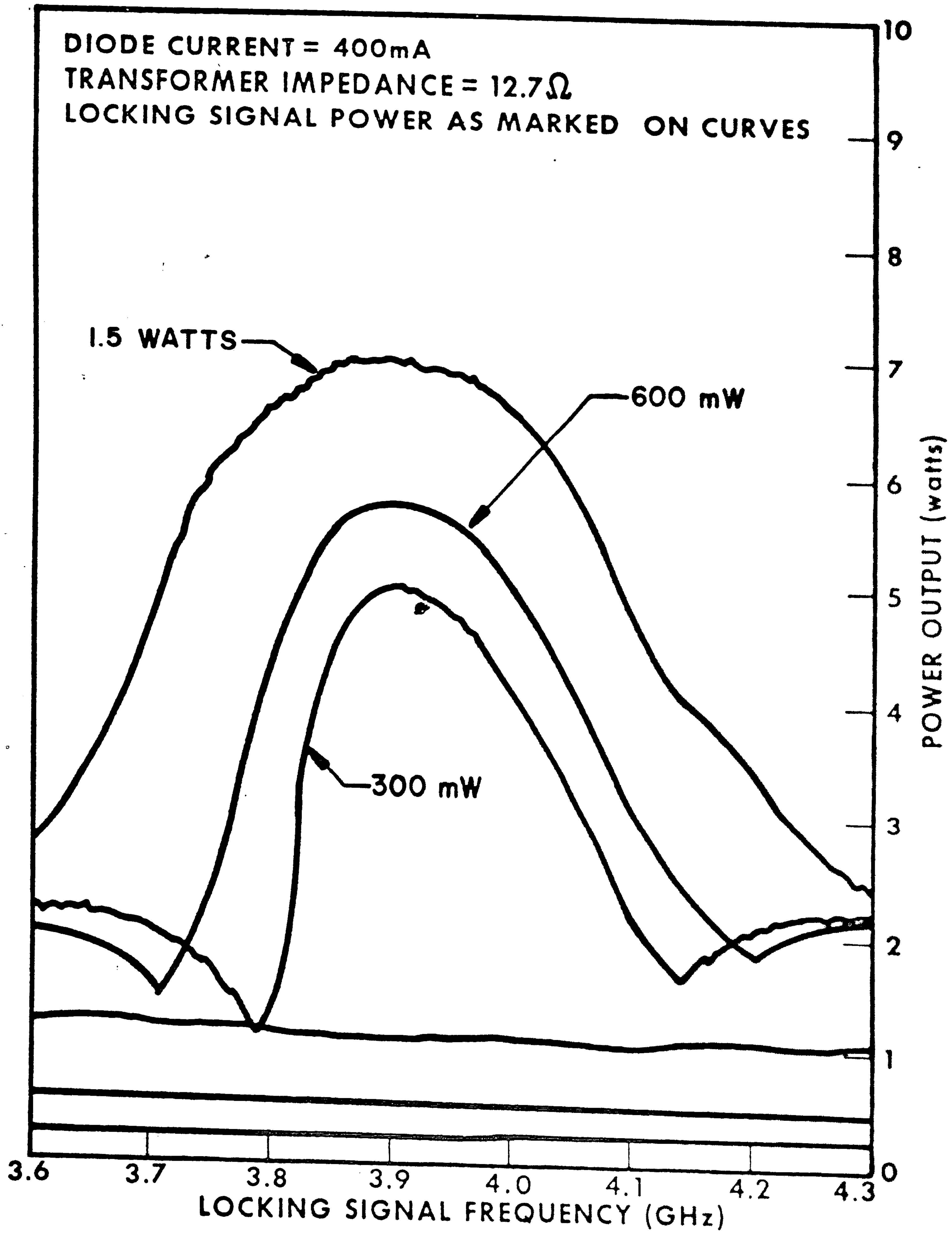
FIGURE 24



LOCKING RANGE FOR

DIODE: # 1

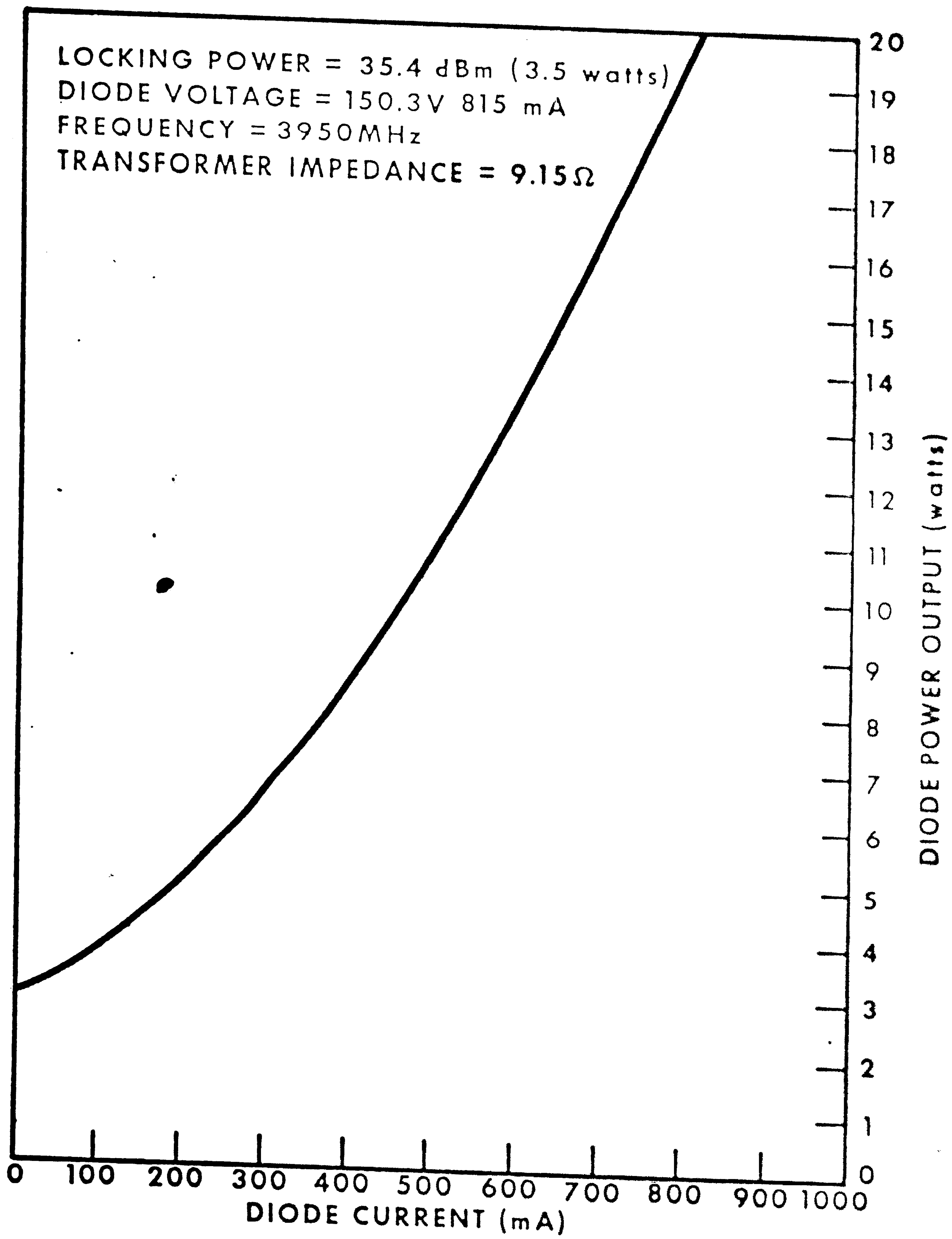
FIGURE 25



LOCKING RANGE FOR DIODE PAIR

DIODES: # 1 AND 2

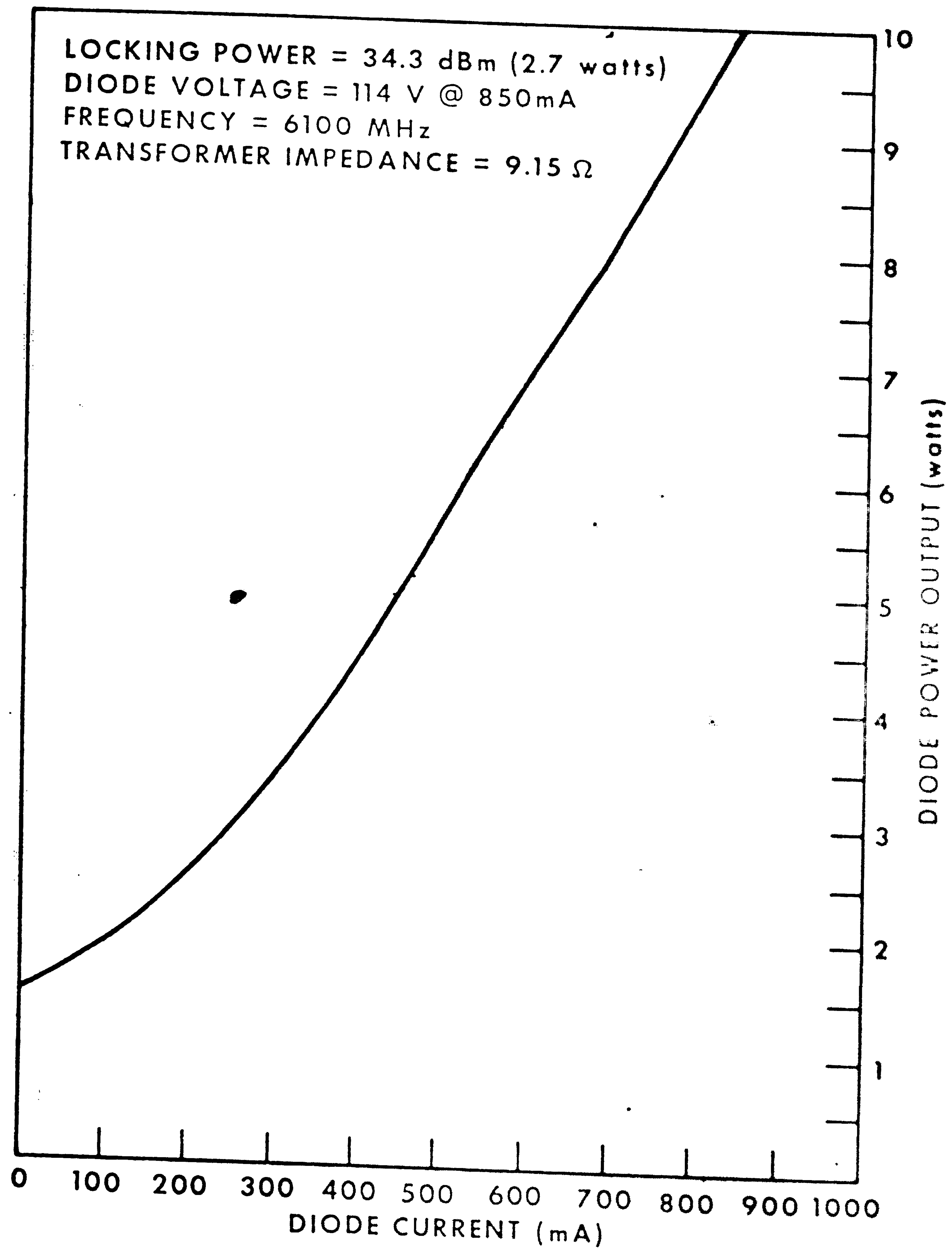
FIGURE 26



DIODE TRIPLET POWER OUTPUT vs. CURRENT

4 GHz

FIGURE 27



DIODE TRIPLET POWER OUTPUT vs. CURRENT

6 GHz AMPLIFIER

FIGURE 28

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